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SOVIET INSTRUMENTATION AND CONTROL TRANSLATION SERIES

Measurement Techniques

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Measurement Techniques

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(Russian Original Dated August, 1961)

1961, No. 8, pp. 601-686 February, 1962

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LINEAR MEASUREMENTS

UNIVERSAL MEASURING COMPARATOR TYPE IKU-2

N. F. Delyunov, E. I. Rozenberg, and E. P. Smirnova

Translated from Izmeritel'naya Tekhnika, No. 8,
pp. 1-3, August, 1961

In 1959 the State opticomechanical plant developed a new universal measuring comparator, model (horizontal length gauge) type IKU-2 [1]. The instrument is intended for direct and differential measurements of internal and external linear dimensions of various articles. The differential dimensions are measured by comparing the measured article with block gauges, templates or reference details. It is intended to start mass production of these instruments in 1961 after they have passed their state testing.

Length measurement range, mm	0-500
Diameter measurement range, mm	
external	0-225
internal	13.5-150
Minimum calibration of the readout device, mm	0.001
Linear scale calibration, mm	1.0
Linear scale measurement range, mm	0-100
Micron scale measurement range, mm	±0.1

Maximum error of this instrument in measuring block gauges (using instrument scale certificate correction), mm: $\pm\left(0.001+\frac{L}{200{,}000}\right)$, where L is the measured length, mm.

Maximum weight of the measured detail, kg	10.0
Over-all dimensions, mm	1150 × 460 × 520
Weight, kg	150

The optical arrangement of the instrument is shown in Fig. 1. Lamp 1 illuminates millimeter scale 10 through prism 2, condenser 3, heat filter 4, filter 5, mirrors 14 and 6, prism 7 and lens 8. The images of the scale graduations are projected by objective 11 through prism 15 onto the plane of the tenths of a millimeter scale 16, which is drawn in the form of double graduations. The images of the millimeter scale graduations and those of the tenths of a millimeter scale are projected onto screen 9 through prism 17, objective 19 and mirrors 13 and 27.

Lamp 1 illuminates grid 23 placed in the focal plane of objective 24 through prism 31, condenser 30, heat filter 29 and filter 28. The grid consists of a plane-parallel plate with a micron scale on one half of it and an index on the other. The micron scale is similar to that of the optimeter tube scale and has 200 calibrations placed symmetrically either side of zero (100 calibrations on each side).

The rays of light are projected in a parallel beam from grid 23 through prisms 18 and 26, objective 24 and prism 22 onto mirror 25 and are reflected from it back onto prism 22, objective 24, prism 26 and 18. The micron scale image is thus produced in the index plane and projected tobether with the index onto screen 9 by means of objective 20 and mirrors 21 and 27.

Mirror 25 is coupled to a vibrating system similar to that used in the optimeter tube. An axial displacement of the measuring stock tip deflects mirror 25 through a certain angle, thus making the image of the micron scale move on the screen with respect to the index. The mechanical and optical parameters of the instrument are chosen in such a way that the displacement of the micron scale equal to one division corresponds to an axial displacement of the measuring tip equal to 0.001 mm.

The graduations of the millimeter scale and the double graduations of the tenths of a millimeter scale appear in the upper part of the screen, whereas the index and the micron scale appear in its lower part. In comparison measurements reading are taken from the micron scale. In direct measurements hundredths and thousandths of a millimeter are read off the micron scale, complete millimeters off the linear scale, and tenths of a millimeter of the tenths of a millimeter scale with its double graduations. The reading shown in the figure amounts to 4.5727 mm.

For centering the lamp, lens 12 is used and the lamp filament is projected onto the screen instead of the millimeter and the tenths of a millimeter scales.

Comparator IKU-2 is shown in Fig. 2. The instrument is mounted on a cast base 1 with strengthening ribs. The top of the base carries flat guides for mounting and directing the displacement of tail spindle stock 2 and measuring stock 5. The middle part of the base is fitted with an objective table 11.

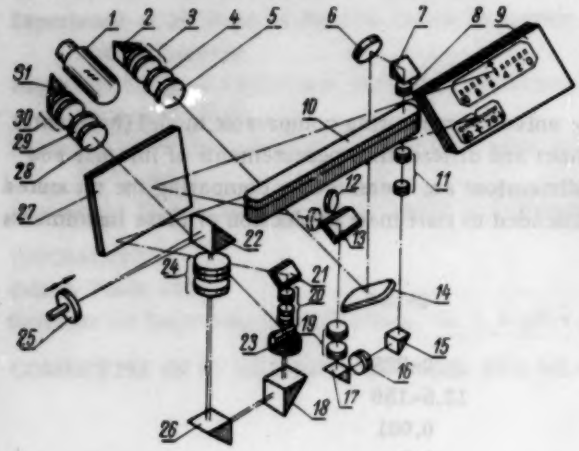

Fig. 1.

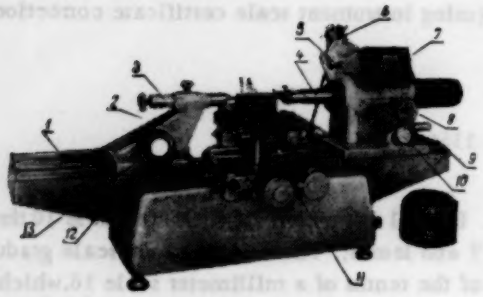

Fig. 2.

The tail spindle stock is displaced along the guides by knob 13 and secured in the required position by lock screw 12. The upper part of the stock carries, along the measuring line of the instrument, tail spindle 3 whose construction is similar to that of the spindle used in the horizontal optimeter IKG [2].

The measuring stock is displaced manually along the guides and secured in the required position by screw 10. The optical system is mounted inside the measuring stock. The forward part of measuring spindle 4 carries a head with a vibrating mirror system coupled to the measuring rod on whose end interchangeable tips are placed. The middle portion of the spindle has a 100-millimeter scale which serves to measure the spindle's displacement. The rough displacement of the spindle is made by means of knob 8, and the accurate displacement by microscrew 9. For measuring internal dimensions, holders with bows are fitted onto the head and tail stock spindles.

The upper part of the measuring stock carries screen 7 and behind it the illuminating system 6. For lighting purposes lamp STs-68 (8 v, 30 w) is used, with a transformer which connects it to the lighting mains (127 v).

The objective table 11 is designed for placing on it and securing articles and various special devices and components to be measured. The mechanism of the table provides all the required movements for displacing the measured object with respect to the line of measurement.

All the attachments produced for set IKG can be used with instrument IKU-2, including vertical centers PP-2, horizontal centers PP-3 and the electrical contact head GK-2 [1]. Specially developed devices include attachments IP-6 and IP-7, which serve for measuring the mean diameter of internal metric threads in rings. By means of these attachments it is possible to measure mean diameters between 17 and 150 mm with a pitch of 1.5-3.5 mm. The principle of measurement involved is similar to that used in attachment IP-4 of optimeter IKG [2].

The expected maximum error in direct measurements of external dimensions by means of this instrument consists of the following errors: the attestation error Δl_1 of the linear scale, the reading error Δl_2 of the micron scale, the error Δl_3 due to the setting of the readout device to normal magnification, the error Δl_4 due to the placing of the millimeter scale graduation between the double graduations of the tenths of a millimeter scale, and the temperature error Δl_5. The total error is determined from the formula $\Delta l = \pm \sqrt{\sum_{i=1}^{i=5} \Delta l_i^2}$. Errors whose value does not exceed 0.2 μ are not included in determining the error Δl.

Instrument IKU-2 operates on the comparator principle. Hence, the comparator error, which is determined by the relation $L(\varphi^2/2)$ (L is the measured length, φ is the slope angle of the measuring spindle), amounts to a small quantity. For L = 100 mm and $\varphi = 10''$ the comparator error does not exceed 0.1 μ. The permissible deviation from the normal of the distance between any two millimeter scale graduations does not exceed $\pm [0.001 + (L/200{,}000)]$ mm, where L is the distance between the graduations.

Since the actual deviation of each graduation from the normal is entered in the certificate attached to the scale and is taken care of by appropriate corrections, the errors in the scale calibration do not affect the instrument measurement errors. It is only necessary to take into account the error of scale attestation, i.e., the errors of the corrections listed in the certificate. The scale of the IKU-2 instrument is certified by means of a stereocomparator as a 2nd grade scale with a maximum error of ± 0.5 μ according to the Committee's Instruction 82-56. Since in measurements, readings are taken of two different graduations of the scale, the maximum error introduced by the millimeter scale amounts to $\Delta l_1 = \pm 0.5\sqrt{2} = \pm 0.7\ \mu$.

The tolerance of the IKU-2 instrument's micron scale readings amounts to ±0.3μ. The error of measurement introduced by the micron scale with readings taken from two different calibrations amounts to $\Delta l_2 = \pm 0.3\sqrt{2} = \pm 0.4\ \mu$.

When the readout system is set for normal magnification, the millimeter scale graduation images are adjusted to fall between the double graduations of the zero and ten divisions on the tenths of a millimeter scale. A setting error of ± 0.5 μ is then allowed. The measurement error due to the setting of the readout device to normal magnification thus amounts to $\Delta l_3 = \pm 0.5\ \mu$.

The error due to the setting of the millimeter scale calibration image between the double calibrations of the tenths of a millimeter scale and the readout error of the micron scale were determined as a combined error when the instrument was checked and measurement results processed. This maximum combined error amounted to ± 0.36 μ. Since in each measurement the graduation image is set at two different points of the tenths of a millimeter scale, the above error enters in the measurement result twice, and its maximum value amounts to $\Delta l_4 = \pm 0.36\sqrt{2} = \pm 0.5\ \mu$.

Errors due to temperature deviations from the normal (20°C), at which measurements should be made, are produced by the incomplete equalization of the temperatures of the measured object and the instrument, and the difference in the linear temperature coefficients of the measured object and the instrument's millimeter scale.

In measurements by means of the IKU-2 set, a difference in temperature of ±0.5°C between the measured object and the instrument scale is allowed. The measurement error for a linear temperature coefficient of steel of $\alpha = 11.5 \cdot 10^{-6}\ \text{deg}^{-1}$ due to this cause amounts to $\Delta l'_5 = \pm 0.5 \cdot 11.5 \cdot 10^{-6} \cdot L\ \text{mm} = \pm 5.75 \cdot 10^{-3} L\ \mu$, where L is the measured length, mm.

For calculation purposes it is possible to assume that the difference in the linear temperature coefficients of the measured object and the instrument scale amounts to $1.4 \cdot 10^{-6}\ \text{deg}^{-1}$. When the temperature of the measured object and the instrument scale deviate from the normal by ± 2°C, a measurement error is produced due to the difference in the linear temperature coefficient of the measured object and the instrument scale which amounts to $\Delta l''_5 = \pm 1.4 \cdot 10^{-6} \cdot 2\ L\ \text{mm} = \pm 2.8 \cdot 10^{-3} L\ \mu$.

Thus, the measurement error due to temperature errors will amount to

$$\pm\sqrt{\Delta l_5'^2 + \Delta l_5''^2} = \pm\sqrt{(5.75 \cdot 10^{-3}L)^2 + (2.8 \cdot 10^{-3}L)^2} = \pm 6.4 \cdot 10^{-3} L\ \mu.$$

The expected maximum error of instrument IKU-2 in direct measurements of external dimensions amounts to

$$\Delta l = \pm\sqrt{0.7^2 + 0.4^2 + 0{,}5^2 + 0.5^2 + (6.4 \cdot 10^{-3}L)^2} = \pm\sqrt{1.15 + (6.4 \cdot 10^{-3}L)^2} \approx \pm\left(1 + \frac{L}{200}\right)\mu,$$

where L is the measured length, mm.

The testing of experimental models of instrument IKU-2 confirmed that the maximum error in direct measurements of external dimensions by means of this instrument does not exceed the expected error obtained by calculations.

Instrument IKU-2 will have a wide field of application in the engineering and instrument-making industry, in various test laboratories and scientific-research institutes.

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2. A. K. Kutai, Handbook on Engineering Production Control [in Russian] (Mashgiz, Moscow, 1956).

PRESSURE STABILIZERS FOR PNEUMATIC INSTRUMENTS USED IN LINEAR MEASUREMENTS

A. P. Kurochkin and F. V. Tsidulko

Translated from Izmeritel'naya Tekhnika, No. 8,
pp. 4-7, August, 1961

The Interchangeability Bureau of the State Committee of Automation and Engineering has carried out an experimental and analytical investigation of air pressure stabilizers for instruments used in checking linear dimensions.

These stabilizers work in the range of input ("mains") pressures of 3-6 A_s, output ("working") pressures of 0.3-2.8 A_s and flows of 5-150 nl/min (sometimes up to 250 nl/min). A peculiarity of these stabilizers consists in stringent requirements for maintaining a constant working pressure. The known designs of pressure stabilizers can be reduced to the four following types of proportional regulators: reversed action (Fig. 1a), direct action (Fig. 1b), amplified reversed action (Fig. 1c), and another version of a direct action regulator (Fig. 1d).

The balance equation for the moving system of a stabilizer shown in Fig. 1a can be represented without taking into account the weight of the system and stiffness of the diaphragm by the following relation of the working pressure to the mains pressure and the air flow:

$$H=\left(\frac{P_1-P_2}{F_{eff}-f_v}\right)-\left(\frac{f_v}{F_{eff}-f_v}\right)P_c-\left[\frac{K_1+K_2}{Bd_v\ (F_{eff}-f_v)}\right]\left(\frac{Q}{P_m+1.03}\right),$$

where H is the working pressure, A_s; P_m is the mains pressure, A_s; Q is the air flow through the stabilizer, nl/min; P_1 and P_2 are the pressures of the main and valve springs of the stabilizer with a closed valve, kg-wt, $P_1 \gg P_2$; K_1 and K_2 are the stiffnesses of these springs, kg-wt cm, $K_1 \gg K_2$; F_{eff} is the effective area of the diaphragm, cm^2; d_v is the diameter of the valve opening, cm; $f_v = \pi d_v^2/4$ is the area of the valve opening, cm^2; $B = 62.3 \cdot 10^3 \cdot \frac{\alpha}{\sqrt{273+t^0}} \cdot \frac{1}{\text{kg-wt} \cdot \text{min}}$ for a supercritical condition of air flow through the valve, i.e., for $\frac{H+1.03}{P_m+1.03} \leq 0.528$, and

$$B=241\cdot10^3\cdot\frac{\alpha}{\sqrt{273+t^0}}\sqrt{\left(\frac{H+1.03}{P_m+1.03}\right)^{1.43}-\left(\frac{H+1.03}{P_m+1.03}\right)^{1.71}}\cdot\frac{l}{\text{kg-wt}\cdot\text{min}}$$

for a subcritical condition of airflow through the valve, i.e. for $\frac{H+1.03}{P_m+1.03} > 0.528$, where α is the flow factor of the valve; t^0 is the air temperature at the stabilizer input.

Let us now examine the relation between H and P_m. For the case when $Q = 0$ the relation of the working pressure to the main pressure is represented by a broken line OAB (Fig. 2). In this graph the ordinates do not represent mains pressure, but the ratio of this pressure to a constant quantity H_{nom}, which is the nominal value of the working pressure equal to an arbitrary value of H in a range of P_m, which is suitable for operation, i.e., where the effect of P_m on H is small. We find that $H = P_m$ in section OA, and in section AB the relation between H and P_m is represented by (1) without a third term in its right-hand side, since $Q = 0$. Moreover (1) becomes linear and the slope angle of the straight line to the abscissa is $\alpha = \tan^{-1}\frac{f_v}{F_{eff}-f_v}$. In stabilizers shown in Fig. 1a it is normal to have $f_v \ll F_{eff}$;

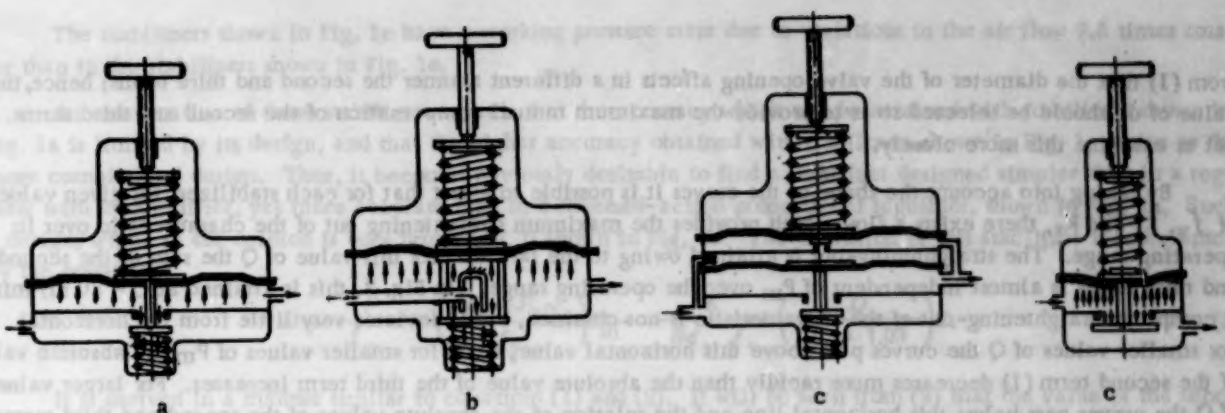

Fig. 1.

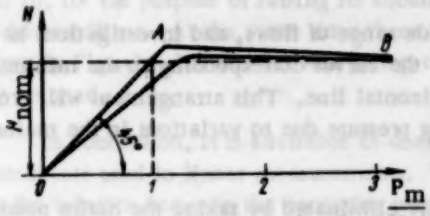

Fig. 2.

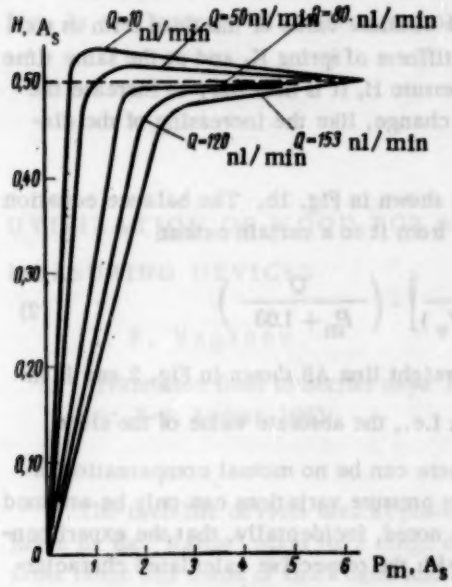

Fig. 3

hence $\frac{f_v}{F_{eff} - f_v} \ll 1$ and, therefore, in section AB the variations of H are considerably smaller than those of P_m which have produced the former. Let us note that $\frac{f_v}{F_{eff} - f_v} \approx \frac{f_v}{F_{eff}}$, since $f_v \ll F_{eff}$.

For $Q \neq 0$, each value of Q corresponds to a curve which always passes below the line OAB. Figure 2 shows an example of such a curve. For an amplification of P_m / H_{nom} these curves approach asymptotically the line AB. If the shape of the curves is examined for a P_m decreasing from its maximum value, it will be seen that the curves deviate increasingly from the straight line AB, and that for small flows the curves deviate but little from AB. Moreover, these curves first rise, reach a maximum and then begin to drop, and for large flows they deviate considerably from line AB and from the start begin to drop. This is quite natural, since the value of the third term in the equation is proportional to the flow Q, and the greater the Q the greater will be the deviation of the characteristic from the straight line AB. In each curve for a certain value of P_m the value of H begins to drop rapidly. Hence, if the stabilizer operates at values of P_m smaller than the above the error in H rises rapidly. The value of P_m for which the curve begins to drop rapidly depends on Q. The greater the Q the greater will be that value of P_m. It will obviously always be greater than H_{nom}. In a correctly designed stabilizer the curve which corresponds to the maximum value of Q does not tend to drop rapidly for the smallest possible value of P_m.

Experimental characteristic curves of stabilizers shown in Fig. 1a and obtained for various values of H and Q agree well with the calculated data.

Figure 3 shows a family of experimental characteristics with respect to mains pressure obtained for one stabilizer. The general shape of curves of stabilizers shown in Fig. 1a is the same for all the stabilizers and under all the operating conditions.

In view of the above it becomes possible to consider the problem of reducing the effect of P_m on H, in other words of "straightening out" the characteristics, i.e., approaching them to a horizontal straight line. It will be seen from (1) that this can be achieved by firstly decreasing the absolute value of the second and third terms in the right-hand side of the equation, and secondly by partly compensating these terms. The absolute values of both these terms are reduced by increasing F_{eff}, i.e., enlarging the inside diameter of the diaphragm. This diameter normally amounts, in stabilizers shown in Fig. 1a, to 60-100 mm and determines the diameter dimension of the stabilizers. A further increase in the internal diaphragm diameter can be made by raising the dimensions of the stabilizer. It will be seen

from (1) that the diameter of the valve opening affects in a different manner the second and third terms; hence, the value of d_v should be selected so as to provide the maximum mutual compensation of the second and third terms. Let us examine this more closely.

By taking into account the shape of the curves it is possible to assert that for each stabilizer, for given values of f_v, K_1 and K_2, there exists a flow which provides the maximum straightening out of the characteristic over its operating range. The straightening-out is attained owing to the fact that for this value of Q the sum of the second and third terms is almost independent of P_m over the operating range. In Fig. 3, this is attained at Q = 80 nl/min. A complete straightening-out of the characteristic is not obtained, but it deviates very little from the horizontal. For smaller values of Q the curves pass above this horizontal value, since for smaller values of P_m the absolute value of the second term (1) decreases more rapidly than the absolute value of the third term increases. For larger values of Q the curves pass below this horizontal line and the relation of the absolute values of the second and third terms is reversed. It is obvious that with a decreasing f_v and, hence, a decreasing f_v/F_{eff}, the straight line AB (Fig. 2) will become less steep, and with a rise in the above values it will become steeper. This effect will turn the family of characteristic curves obtained for different flows in a direction above or below the horizontal.

Stabilizers used in pneumatic measurements operate over a very wide range of flows, and investigations have shown that the value of f_v should be chosen for each range so as to place the curves corresponding to the minimum and maximum flows approximately symmetrically with respect to the horizontal line. This arrangement will provide in the operating range the smallest possible maximum error of the working pressure due to variations in the mains pressure.

It is interesting to note that the second term in (1) can be completely eliminated by taking the mains pressure off the valve. There exist several design versions for achieving this effect. However, in stabilizers used in pneumatic measurements this would only lead to an increased error due to variations in the mains pressure, since the mutual compensation of the second and third terms would then be disturbed.

A reduction in the stiffness of the mains spring K_1 would decrease the absolute value of the third term as well as the error of the stabilizer as a whole. However, in order to reduce the stiffness of spring K_1 and at the same time preserving the effort P_1 provided by it, which corresponds to the working pressure H, it is necessary to increase the length of the spring, and hence the height of the stabilizer. However, this change, like the increasing of the diaphragm diameter, is only expedient within certain limits.

Let us now examine the relation between H and P_m for the stabilizer shown in Fig. 1b. The balance equation for this stabilizer derived with the same assumptions as equation (1) differs from it to a certain extent:

$$H = \left(\frac{P_1 - P_2}{F_{eff} + f_v}\right) + \left(\frac{f_v}{F_{eff} + f_v}\right) P_m - \left[\frac{K_1 + K_2}{Bd_v(F_{eff} + f_v)}\right] \cdot \left(\frac{Q}{P_m + 1.03}\right). \qquad (2)$$

For Q = 0, straight line AB, plotted from this equation, is similar to straight line AB shown in Fig. 2 and is inclined to the x-axis at an angle of $\alpha = \tan^{-1} \cdot \frac{f_v}{F_{eff} + f_v} \approx \tan^{-1} \cdot \frac{f_v}{F_{eff}}$; i.e., the absolute value of the slope angle is virtually the same as in (1), but its sign is reversed. In this case there can be no mutual compensation of the second and third terms. Hence the operating pressure error due to mains pressure variations can only be attained by decreasing the value of the second and third terms (2). It should also be noted, incidentally, that the experimental characteristics $H = f(P_m)$ of this stabilizer are also in good agreement with the respective calculated characteristics.

Let us now examine the relation between the working pressure H and the air flow Q. For stabilizers shown in Figs. 1a and 1b we find from (1) and (2) that H is a linear function of Q, and that with a rising Q the working pressure decreases. In order to reduce the effect of Q on H it is advisable to increase F_{eff} and d_v, and to decrease K_1. However, it has already been pointed out that a rise in F_{eff} and a drop in K_1 lead to increased dimensions of the stabilizer, and that increasing d_v above a certain optimum value leads to an increased error due to variations of the mains pressure. On the basis of experimental data the following conclusions can be drawn: the stabilizer shown in Fig. 1b is approximately equivalent to stabilizer shown in Fig. 1a with respect to its complexity and dimensions, but it is less accurate. Therefore it is not advisable to use stabilizer shown in Fig. 1b for pneumatic testing.

The stabilizers shown in Fig. 1c have a working pressure error due to variations in the air flow 2.5 times smaller than in the stabilizers shown in Fig. 1a.

An examination of these stabilizers reveals that the accuracy which can be attained with stabilizers shown in Fig. 1a is limited by its design, and that the higher accuracy obtained with stabilizers shown in Fig. 1c is due to their more complicated design. Thus, it becomes obviously desirable to find a stabilizer designed simpler than in a regulator with an amplifier, yet more accurate than in a reversed-action proportional regulator, shown in Fig. 1a. Such a design, which in our opinion is very promising, is shown in Fig. 1d. The operation of this stabilizer is represented by the equation:

$$H = \frac{P}{f_v} + \left(\frac{f_v - F_{eff}}{f_v}\right)P_m - \frac{K}{Bd_v f_v}\left(\frac{Q}{P_m + 1.03}\right)E. \qquad (3)$$

It is derived in a manner similar to equations (1) and (2). It will be seen from (3) that the value of the second term in the right-hand side of the equation can be reduced virtually to zero by making $F_{eff} = f_v$, and the value of the third term is several times smaller than that in stabilizers shown in Figs. 1a and 1b, owing to the considerably larger value of d_v. In this stabilizer there is no necessity to increase F_{eff}, as it is done in the design shown in Figs. 1a and 1b, for the purpose of raising its accuracy of operation. Thus, it becomes possible to considerably reduce the size of the stabilizer. At the same time there is no longer any need to reduce d_v, as it has to be done in the stabilizers shown in Figs. 1a, 1b, and 1c, and owing to a large d_v the stabilizer error due to variations in the flow is considerably reduced.

In conclusion, it is advisable to deal with the application of various air pressure stabilizers for instruments and transducers used in linear measurements. The special work carried out by the Interchangeability Bureau in establishing the relation between the operating pressure error, and hence, the error of measurement, has made it possible to specify the requirements for various measuring systems with respect to pressure stabilizer accuracy. Knowing the actual accuracy characteristics of stabilizers, it has been made possible to specify their ranges.

The basic conclusions arrived at in this work with respect to stabilizers show that stabilizers made according to Figs. 1a and 1c cover all the requirements of pneumatic measuring devices in different spheres of application.

UTILIZATION OF WOOD FOR MAKING LARGE-SIZE MEASURING DEVICES

I. P. Vaganov

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The metallic devices used at present for measuring large sizes are very heavy. Thus, the steel micrometer made by the "Kalibr" plant for a range of 1750-2000 mm weights 20 kg, diameter-measuring caliper gauges for the same range and made of steel tubes and tubular caliper gauges for linear measurements up to 6000 mm weigh 25 kg and more, and slide gauges measuring up to 4000 mm weigh 30 kg. At the same time the accuracy of the above devices is low owing to their low rigidity and large temperature errors. They can be used under the most favorable conditions for measuring articles of the grade of accuracy of 3a-5 GOST (All-Union State Standard) 2689-54.

Plastic gauges including Fiberglas-reinforced epoxy resin gauges are used in the USA. Normally they are cut out or pressed out of plates, and their sizes are small [1, 2]. The effect of temperature on these gauges or their error characteristics have not been published anywhere.

Attempts to reduce the weight of diameter-measuring and linear caliper gauges by making them of aluminum alloys have not produced any positive results either with respect to their rigidity or their accuracy.

Attempts have been made in the Soviet Union to manufacture large gauges of laminated wood plastics (DSP). Preliminary tests of a 1200 mm micrometer made of a laminated wood plastic provided satisfactory results [3]. Therefore, in the last two decades the idea of the possible use of wood in making gauges for measuring large dimensions has been gaining ground. They are expected to be of a higher precision than metal gauges.

The well-known Czechoslovak scientist Professor N. N. Savin asserts [4] that insulation of the wood from humidity, a rational design of gauges, and the application of the basic rules of measurement technology to wood gauges will make them serviceable and completely reliable for ensuring interchangeability of large-size machine parts. The same opinion about wood as a material suitable for gauges is held by a number of other experts.

The physicomechanical properties of gauges will vary with the type of wood used, but these variations are insignificant as compared with similar differences in grades of tool steel. In view of the above, we ascribe, in any further considerations, certain mean properties to wood irrespective of its type (pine, spruce, fir), taken with a relative moisture content of 12%. With these assumptions we have arrived at the following conclusions:

1) the specific gravity of soft wood is 18 times lower than that of steel;

2) the thermal conductivity coefficient of wood is up to 100 times lower (along the grain) and up to 350 times lower (across the grain) than that of steel, and it is only a few times (2 to 6) larger than that of the best lagging materials (slag, cork plates, mineral wadding, felt, etc.);

3) the elasticity modulus of wood is 20 times lower than that of steel (along the grain);

4) the permissible tensile and compressive stresses in steel are 20 times higher than those in wood; however, these stresses per unit weight are about the same in wood and steel.

By comparing the properties of woods of various types, [4, 5] it was found that fir was the most suitable for making gauges. Of all the coniferous species fir has the lowest specific gravity (its volumetric weight is equal on an average to 0.38 g/cm^3), it is more homogeneous, it dries relatively quickly in fresh air and tends to split less than other soft woods; moreover it is less knotty. An important defect of wood, which until recently has limited its use for gauges, consists in considerable variations in its linear dimensions with the humidity of the surrounding air. For a relative humidity of 1% the linear dimensions change in the tangential direction by 0.16% for fir, by 0.24% for spruce and by 0.32% for oak. Considering that in the mean latitudes of our country the air temperature varies during a year from −35 to + 35°C, and relative humidity from 40 (May) to 90% (December), the final moisture content of an unprotected air-dried wood may attain anything from 7 to 23% [5].

Special experiments were conducted [4, 6] for determining the effect of moisture on unprotected (without covering) hole gauges made of fir and maple boards with the best yearly ring structure both in the transverse and longitudinal directions. These boards were previously dried in stacks for 1.5-2 years in the open air. These data were processed by us and are shown in a diagrammatic form (see Fig.) where the observation time t in weeks is plotted along the x-axis, and the deviations in the hole gauge length l in μ, and the relative humidity ε in % along the y-axis. The data of three-meter hole gauges is referred to 1 m.

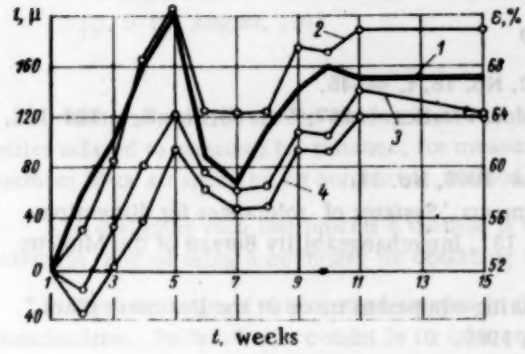
It will be seen from the diagram that curve 2 showing the variations in the meter hole gauge length almost completely coincides with curve 1 showing the variations of the relative humidity of air; however, when the relative humidity after a lapse of 5 weeks fell from 72 to 60%, the variations in the hole gauge length were slowed down and henceforth followed with a lag the variations of air humidity.

The deviations in the length of the 3-meter hole gauge No. 3 (curve 3) are smaller than those of the meter hole gauge on an average by 45%. The deviations in the length of the 3-meter hole gauge No. 4 (curve 4) are in turn smaller than the variations of hole gauge No. 3 on an average by 30%, due to the effect of glue which acts as a hydrophobic substance preventing the penetration of moisture from the air into the wood. A similar effect is observed in hole gauges made of maple. The numerical values in the latter case almost completely coincide with those obtained for pine hole gauges with very small individual deviations on the lower side.

We carried out special experiments with wood half-meter hole gauges of a rectangular cross section made for this purpose at the experimental workshop of the Sverdlovsk branch of the VNIIM (All-Union Scientific-Research Institute of Metrology) from fir boards which had been dried naturally for two years. These experiments have shown that for a relative air humidity variation of 1% the meter models changed their dimension along the grain by 10 μ and across the grain by 100 μ. In samples which were glued "crosswise" from two or three boards at an angle of 45° to the

axis of the sample (the direction of the grain in one of the boards was at 90° to that in the other), the length changed by 16 μ; moreover, the size of the samples begins to change with a lag with respect to the air humidity variations (in some cases the lag amounted to two days).

In order to raise the hydrophobic properties of wood there exist many methods of covering or impregnating it [5]. Comparative tests [4] of moisture absorption by wood covered with various lacquers and paints and then placed in a humid atmosphere (about 100% relative humidity) have shown that if the absorption of a completely nonhygroscopic material is taken to be 0%, the absorption of wood covered with aluminum sheets (Foil) over asphalt or shoe varnish amounts to 0-2%; when covered with three layers of aluminum powder in turpentine paints or over an asphalt base it is 2%; with three layers of aluminum powder over shoe varnish, or aluminum foil over celluloid lacquer or oil paints it is 6-7%; with a consistent thick layer of paraffin it is 9%; with three layers of asphalt, rubber or enamel carriage lacquer it is 10-12%; with three layers of shellac, or cellulose enamel or varnish it is 13-27%; with oil paint or three layers of white lead or zinc oxide it is 30%; with three layers of graphite paint or shoe varnish it is 40%; with linseed oil or drying oil it is 80-82%; with furniture wax it is 90-92%, and unprotected wood is 100%.



Variations in the length of a wooden hole gauge with respect to the relative humidity of air. 1) Relative humidity of air, %; 2) variations in the length of a meter hole gauge No. 1 with a cross section of 20 × 35 mm made from a solid fir board; 3) variations in the length of a 3-meter hole gauge No. 3 with a cross section of 22 × × 60 mm (in the middle of 22 × 80 mm) made of a solid fir board; 4) variations in the length of a 3-meter hole gauge No. 4 of the same cross section made from four solid fir boards (glued together).

The Uralmash plant made experimental samples of wood (fir) indicating diameter-measuring and linear caliper gauges for different measurement ranges. In order to raise the rigidity and strength of the gauges the walls of their frames were glued from separate layers 8 mm thick with the grain of each layer at 90° to the preceding and at 45° to the mean line of the gauge frame. Experiments with fir samples have shown that such a position of the grain is most efficient and provides the least deformation. In order to reduce the weight of the gauges they were made hollow with an internal tensioning member at an angle to the axis of the frame. The gauges were covered by five layers of perchloro-vinyl enamel type PKhV-715.

In the experimental determination of the effect of humidity on the set dimensions of a caliper gauge a shaft model (of an appropriate diameter) was measured in the same position, namely, when it was placed on a marking-off (test) table and the gauge was in the hands of the operator. The air temperature in the room was within the limits of 20.2 to 21.3°C. The temperature variations did not exceed 0.15°C per hour and 0.5°C per shift. Air humidity varied in the range of 35 to 55%. The root-mean-square error of a series of measurements in determining dimensions by means of the diameter-measuring caliper gauge with a range of 2000-2250 mm amounted to ± 6.5 μ for a frame height of 350 mm, and to ± 9.3 μ for a frame height of 500 mm. This indicates that the dimensions were determined within the accuracy of a certificate obtained for the model on an end measuring machine.

In determining the mean value of a set gauge dimension the quadratic mean error of the result was found to equal ± 2.9 μ for a gauge with a frame height of 350 mm and to ± 6.9 μ for a gauge with a frame height of 500 mm, thus indicating a good agreement of the results (especially if it is taken into consideration that the mean value of the set dimension was obtained with considerable variations in air humidity). Thus the dimension of a humidity-protected gauge is virtually (within the accuracy of measurements) independent of air humidity.

Moreover, the tests have shown that the rigidity of wood diameter-measuring caliper gauges in the range of 1750-2000 mm is considerably greater than that of similar metal gauges of different design, as well as of steel micrometers of the same dimensions. The wood diameter-measuring caliper gauge of 2000-2250 mm and a frame height of 500 mm was found to be particularly rigid.

SUMMARY

Wood diameter-measuring caliper gauges are convenient to use, provide more efficient and accurate measurements as compared with the best metal gauges and micrometers, and the ambient air temperature virtually has no effect on measurement results.

Wooden measuring equipment has a number of advantages as compared with metal equipment, providing the peculiarities of wood are taken into account. The use of such equipment should raise by 1-2 grades the accuracy in measuring articles according to GOST 2689-54.

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MECHANICAL MEASUREMENTS

CERTAIN PROBLEMS IN DESIGNING INSTRUMENTS WITH ELASTIC SENSING ELEMENTS

A. I. Ioffe

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Instruments with elastic sensing elements are widely used for indirect measurements of various physical quantities related to pressure; for instance, for measuring height by means of the barometric method, determining Mach number M or air speed by an aerometric method, etc.

In designing such instruments a method is widely used based on determining the characteristics of the sensing element and on using a computer for obtaining the best approximation to the required relationships [1, 2, 3].

The advantage of this method consists in the fact that the approximation can be obtained by means of simple mechanisms. Its drawbacks consist in its inherent error and the incomplete application of the possible reduction of the manufacturing error. With an increasing measuring range these defects become more pronounced and, in certain cases, the instruments fail to meet their requirements with respect to accuracy.

In this article we examine instruments designed without inherent errors (in which actual and not approximate relationships are found), as well as the problem of raising production accuracy by using a sensing element with nonlinear characteristics.

Let the measured value (A) be related to pressure (p) by equation

$$A = F(p). \tag{1}$$

Let us assume that F(p) is a monotonically rising or falling function, which in fact is the case for many physical relationships. The instrument for measuring the value of A can be represented by the block schematic shown in Fig. 1.

The transducer is used for the measurement of pressure and its functional conversion into a signal determined by the characteristic of the sensing element $\varphi(p)$ and for the conversion of pressure into a physical quantity which is fed to the computer (displacement, voltage, etc.).

The computer is intended for providing the relationship between the measured quantity and the quantity transmitted by the transducer.

Let us show that the accuracy of the instrument depends on the shape of the sensing element characteristic. We shall also assume that the transducer accuracy does not depend on the shape of the sensing element characteristic. This condition occurs when errors of the sensing element (hysteresis and residual aftereffect) amount to a negligible part of the total transducer error.

The relative instrument error due to the error in measuring pressure is

$$\frac{\Delta A}{A} = \frac{\partial A}{\partial p} \cdot \frac{\Delta p}{A}, \tag{2}$$

where Δp is the absolute error in measuring pressure.

The sensing elements are usually designed with a monotonically rising or falling characteristic which is expressed by one of the following functional relationships:

a linear $W = k_1(p)$ (3)

a logarithmic $W = k_2 \lg p$, (4)

or an exponential $W = k_3 p^n$, (5)

where W is the movement of the sensing element, and $\underline{k}$ is the coefficient of proportionality.

The expressions for the relative error of the transducer can be written:

in the case of the sensing element with a linear characteristic as

$$\delta_1=\frac{\Delta p}{P_{max}-P_{min}}; \tag{6}$$

with a logarithmic characteristic as

$$\delta_2=\frac{\Delta \lg p}{\lg P_{max}-\lg P_{min}}; \tag{7}$$

and with an exponential characteristic as

$$\delta_3=\frac{\Delta p^n}{P^n_{max}-P^n_{min}}, \tag{8}$$

where Δp, $\Delta \lg p$, and Δp^n are the absolute errors in the quantities transmitted by the transducer; P_{min} and P_{max} are the minimum and maximum values of the measured pressures.

Let us determine for each type of transducer the measurement error for pressure Δp.

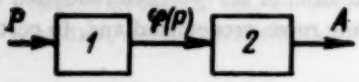

Fig. 1. 1) Transducer; 2) computer.

For transducers with a linear characteristic we have

$$\Delta p=(P_{max}-P_{min})\cdot\delta_1. \tag{9}$$

For those with a logarithmic characteristic:

$$\Delta p=\frac{\partial p}{\partial \lg p}\cdot\Delta\lg p=p(\lg P_{max}-\lg P_{min})\cdot\delta_2. \tag{10}$$

And for those with an exponential characteristic:

$$\Delta p=\frac{\partial p}{\partial p^n}\cdot\Delta p^n=\frac{(P^n_{max}-P^n_{min})}{n\cdot p^{n-1}}\cdot\delta_3. \tag{11}$$

It will be seen from (9), (10), and (11) that Δp depends on the sensing element characteristic.

Having determined the instrument error in relation to the shape of the sensing element characteristic, a sensing element is chosen which provides the smallest error.

If the measured quantity is the function of two or more pressures, the shape of the sensing element characteristic is determined by finding the partial errors due to each pressure and by the above method analyzing all the partial errors and their sum.

According to the shape of the sensing element characteristic the computer performs various mathematical operations. The measurement error due to the accuracy of the computer must not depend on the type of the mathematical problem or must be considerably smaller than the transducer error.

Devices which provide the same accuracy for the solution of various mathematical problems consist, for instance, of electromechanical analogue computers. They are widely used in instrument-making and provide solutions with an error of 0.01-0.1%, which is virtually one order higher than the accuracy of pressure transducers. Digital computers provide solutions with virtually any required accuracy.

As an example let us examine an instrument which measures the Mach number by an aerometric method. The Mach number is related to the dynamic pressure P_d and the static pressure P_{st} by the expressions:

for numbers $M<1$

$$M=2.23\cdot\sqrt{\left(\frac{P_d}{P_{st}}+1\right)^{\frac{1}{3.5}}-1}, \tag{12}$$

for numbers $M>1$

$$\frac{P_d}{P_{st}}=\frac{166.922M^7}{(7M^2-1)^{2.5}}-1. \tag{13}$$

It will be seen from (12) and (13) that number M is a function of the ratio P_d/P_{st}. The relation between number M and P_d/P_{st} is shown in Fig. 2.

Let us examine the following three methods of obtaining number M.

1. Pressures P_d and P_{st} are measured, their ratio is found, and a function generator is used to determine the relationship

$$M=f\left(\frac{P_d}{P_{st}}\right).$$

2. The values of lg P_d and lg P_{st} are determined, the logarithms are subtracted and a function generator is used to solve the relationship:

$$M=f(\lg P_d-\lg P_{st}).$$

3. The relation $M = f(P_d/P_{st})$ is approximated by expression $M = A(P_d/P_{st})^n$, where A and n are certain constant parameters depending on the shape of the curve $M = f(P_d/P_{st})$.

A and n can be determined from two points of a curve (for instance 1 and 2 in Fig. 2).

Then:

$$n=\frac{\lg M_1-\lg M_2}{\lg\left(\frac{P_d}{P_{st}}\right)_1-\lg\left(\frac{P_d}{P_{st}}\right)_2};\quad A=\frac{M_1}{\left(\frac{P_d}{P_{st}}\right)_1^n}$$

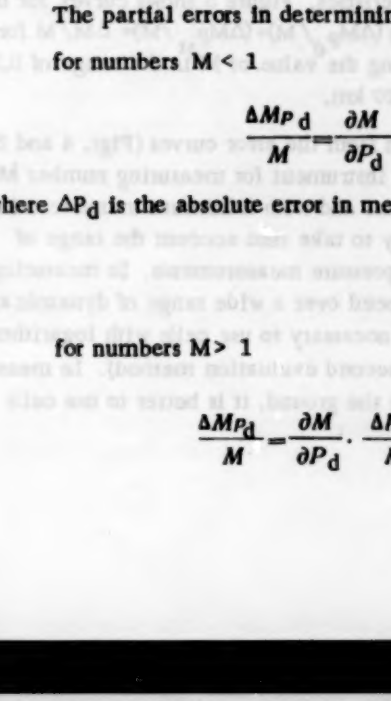

Fig. 2.

Having determined P_d^n and P_{st}^n from the above and obtained the ratio P_d^n/P_{st}^n by means of a function generator which determines the difference between the curves $M = f(P_d/P_{st})$ and $M = A(P_d/P_{st})^n$, we find number M. Let us note that in this instance the function generator has a much smaller deviation from linearity than in the first or second cases.

The block schematics of instruments for each of the above methods of finding number M are shown in Fig. 3.

Pressure-measuring sensing elements made in the form of diaphragm or aneroid cells with nonlinear characteristics make it possible to combine measurements with functional conversion of lg P_d and lg P_{st} for the second method, and P_d^n and P_{st}^n for the third method of determining number M.

In order to raise the accuracy of the instrument, let us utilize the property of the sensing elements to provide functional conversion of the measured quantities according to the required law without increasing the error of the transducer.

The partial errors in determining number M due to the errors in measuring the dynamic and static pressures are:

for numbers M < 1

$$\frac{\Delta M_{P_d}}{M}=\frac{\partial M}{\partial P_d}\cdot\frac{\Delta P_d}{M}=\frac{(M^2+4.9729)^{3.5}-274,22}{7M^2(M^2+4.9729)^{2.5}}\cdot\frac{\Delta P_d}{P_d}=\kappa_1\frac{\Delta P_d}{P_d}. \tag{14}$$

where ΔP_d is the absolute error in measuring the dynamic pressure

$$\kappa_1=\frac{(M^2+4.9729)^{3.5}-274.22}{7M^2(M^2+4.9729)^{2.5}};$$

for numbers M > 1

$$\frac{\Delta M_{P_d}}{M}=\frac{\partial M}{\partial P_d}\cdot\frac{\Delta P_d}{M}=(7M^2-1)\ \frac{[166.92M^7-(7M^2-1)^{2.5}]}{1168.5M^7(2M^2-1)}\cdot\frac{\Delta P_d}{P_d}=\kappa_2\frac{\Delta P_d}{P_d}. \tag{15}$$

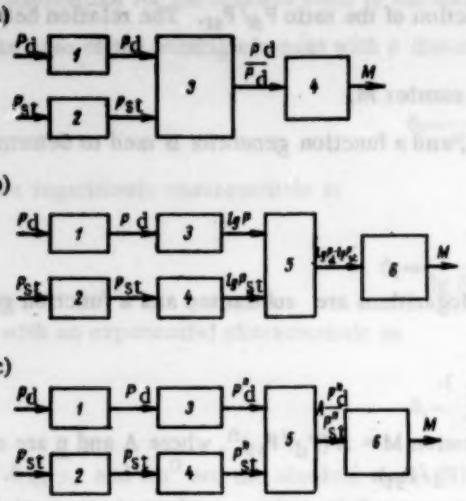

Fig. 3. Block schematic of instruments for determining Number M. a) By the first method: 1) P_d meter; 2) P_{st} meter; 3) divider of P_d/P_{st}; 4) function generator. b) By the second method: 1) P_d meter; 2) P_{st} meter; 3) function generator $P_d \rightarrow \lg P_d$; 4) function generator $M = f(\lg P_d - \lg P_{st})$; c) By the third method: 1) P_d meter; 2) P_{st} meter; 3) function generator $P_d \rightarrow P_d^n$; 4) function generator $P_{st} \rightarrow P_{st}^n$; 5) divider $A \cdot P_d^n/P_{st}^n$; 5) divider $A \cdot P_d^n/P_{st}^n$; g) function generator $M = f[A \cdot (P_d^n/P_{st}^n)]$.

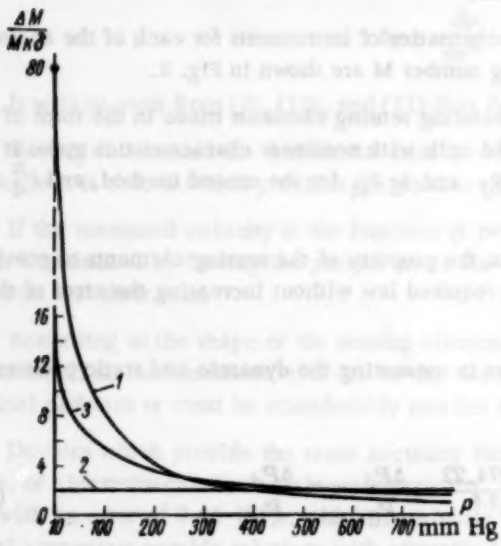

Fig. 4. 1) Transducer with a linear characteristic; 2) transducer with a logarithmic characteristic; 3) transducer with an exponential characteristic.

where

$$\kappa_2 = \frac{(7M^2-1)[166.92M^7-(7M^2-1)^{2.5}]}{1168.5M^7(2M^2-1)};$$

for numbers M < 1

$$\frac{\Delta M_{P_{st}}}{M} = \frac{\partial M}{\partial P_{st}} \cdot \frac{\Delta P_{st}}{M} = \frac{(M^2+4.9729)^{3.5}-274.22}{7M^2(M^2+4.9729)^{2.5}} \times \times \frac{\Delta P_{st}}{P_{st}} = \kappa_1 \frac{\Delta P_{st}}{P_{st}}. \quad (16)$$

where ΔP_{st} is the absolute error in measuring static pressure;

for numbers M > 1

$$\frac{\Delta M_{P_{st}}}{M} = \frac{\partial M}{\partial P_{st}} \cdot \frac{\Delta P_{st}}{M} = (7M^2-1) \times \times \frac{[166.92M^7-(7M^2-1)^{2.5}]}{1168.5M^7(2M^2-1)} \cdot \frac{\Delta P_{st}}{P_{st}} = \kappa_2 \frac{\Delta P_{st}}{P_{st}}. \quad (17)$$

Having determined ΔP_d and ΔP_{st} we find from (9), (10), and (11) the value of partial errors $\Delta M_{P_d}/M$ and $\Delta M_{P_{st}}/M$ for various types of transducers.

The computation results are shown in the table.

It will be seen from the formulas given in the table that the errors in measuring the number M for equal transducer errors (δ) depend on the pressure law of the transducers.

Figure 4 shows the curves of the relation of ΔM_{P_d} per $Mk\delta$ to pressure for transducers with different sensing element characteristics. The pressure measuring range was taken as 810-10 mm Hg.

In plotting the exponential characteristic curve it was assumed that n = 0.5. In fact n may vary according to the range in which number M is measured.

It will be seen from Fig. 4 that at small pressures the errors in measuring number M rise rapidly in transducers with linear characteristics as compared with those having logarithmic characteristics. Figure 5 shows curves for the sum of partial errors $(\Delta M_{P_d}/M)+(\Delta M_{P_{st}}/M)=\Delta M/M$ for instruments measuring the value of M in the range of 0.2-2.6 at heights of 0-20 km.

It will be seen from the error curves (Figs. 4 and 5) that in designing an instrument for measuring number M without inherent errors and with minimum manufacturing errors, it is necessary to take into account the range of dynamic and static pressure measurements. In measuring number M and air speed over a wide range of dynamic and static pressures it is necessary to use cells with logarithmic characteristics (the second evaluation method). In measuring number M and air speed at large dynamic and static pressures, i. e., near the ground, it is better to use cells with linear characteristics (the first evaluation method).

Type of sensing element characteristic	Values of partial errors
Linear	$\frac{\Delta M P_d}{M} = \kappa \frac{(P_{d\max} - P_{d\min})}{P_\partial} \cdot \delta$ $\frac{\Delta M P_{st}}{M} = \kappa \frac{(P_{st\max} - P_{st\min})}{P_{st}} \cdot \delta$
Logarithmic	$\frac{\Delta M P_d}{M} = \kappa(\lg P_{d\max} - \lg P_{d\min}) \cdot \delta$ $\frac{\Delta M P_{st}}{M} = \kappa(\lg P_{st\max} - \lg P_{st\min}) \cdot \delta$
Exponential	$\frac{\Delta M P_d}{M} = \kappa \frac{(P^n_{d\max} - P^n_{d\min})}{nP^n_\partial} \cdot \delta$ $\frac{\Delta M P_{st}}{M} = \kappa \frac{(P^n_{st\max} - P_{st\min})}{nP^n_{st}} \cdot \delta$

$k = k_1$ for numbers $M < 1$, and
$k = k_2$ for numbers $M > 1$.

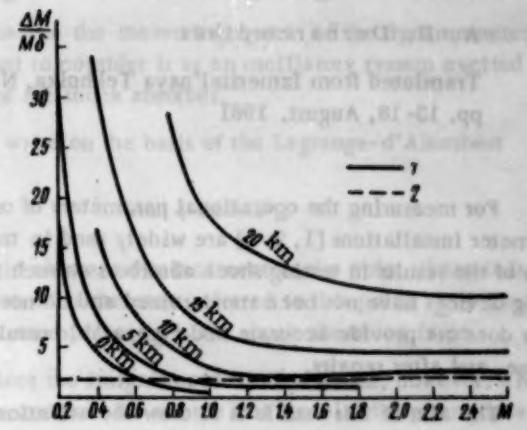

Fig. 5. 1) Errors of an instrument with a linear characteristic; 2) errors of an instrument with a logarithmic characteristic.

The advisability of using cells with exponential characteristics (the third evaluation method) is determined for each specific case according to the value of n and the range of the measured pressures.

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ACCURACY IN MEASURING THE RESISTING FORCE BY MEANS OF ELASTIC ELEMENTS

A. D. Derbaremdiker

Translated from Izmeritel'naya Tekhnika, No. 8, pp. 13-18, August, 1961

For measuring the operational parameters of oscillation dampers in hydraulic shock absorbers,* special dynamometer installations [1, 2, 3] are widely used in transport engineering. However, the problem of a correct evaluation of the results in testing shock absorbers on such installations has not been solved. The parameters of the measuring devices have not been standardized and do not completely meet the requirements specified for similar devices. This does not provide accurate and comparable results in check testing shock absorbers in the manufacturing plants, in use, and after repairs.

The aim of this article is to draw the attention of research and production workers to the size and the characteristic of errors which arise in testing shock absorbers by means of the dynamometer installations with elastic measuring elements, and to indicate methods for evaluating these errors both under laboratory and production conditions. Moreover, the relationships thus obtained provide a reasonable basis for designing new installations and for improving them.

In order to solve the above problems it is necessary to examine the interaction between the measured unit and the measuring device, which arises during the testing of shock absorbers on the dynamometer installation.

The principle of operation of the installation consists in providing the appropriate shock absorber units with an oscillatory movement. This generates in the shock absorber resisting forces which are registered by the elastic measuring element of the installation (Fig. 1).

For a complete oscillatory cycle (one turn of drive 1), shock absorber 3 is subjected consecutively to stretching and compression. The resisting forces which then arise and operate in a direction opposite to the movement of the drive are transmitted through lever 4 to torsion bar 5, which becomes twisted. Lever 6 and its pointer are then deviated from the vertical position either to the right or the left, and pointer pencil 7 draws a line on board 8. Since the board is fixed to slider 2 and is displaced vertically with it, the line drawn by the pencil forms a closed curve (Fig. 2). The deviation of the curve to the right of the zero line indicates in the appropriate scale the effort resisting stretching, and the deviation to the left indicates the effort resisting compression.

The measuring elastic system of the installation is intended for recording the resisting force of the shock absorber with respect to the speed of the piston displacement during oscillations which would exist under normal operations of the shock absorber. Hence the measuring process is dynamic. However, the elastic system is calibrated statically, i.e., the deviation of the dynamometer is determined by suspending reference weights at the end of lever 4.

A certain discrepancy between the static and dynamic effect of the force on the elastic oscillating system, which the above device in fact represents, produces errors due to the inertia of the measuring system.

Another source of errors is provided by the measured unit whose hydraulic resistance affects the measuring system in two ways: on the one hand it produces spurious oscillations, and on the other it takes an active part in suppressing them. Hence, any variation in the resistance parameters of the shock absorber changes the oscillatory parameters of the measuring system, which must also be taken into account for a correct evaluation of the measured unit operating parameters.

Let us now examine these peculiarities in greater detail.

The resisting force developed by the hydraulic shock absorber is in a general form proportional to the speed of vibrations or, to a certain extent, to the speed of liquid flow.

* Hydraulic shock absorbers are at present used in all the suspensions of road transport vehicles.

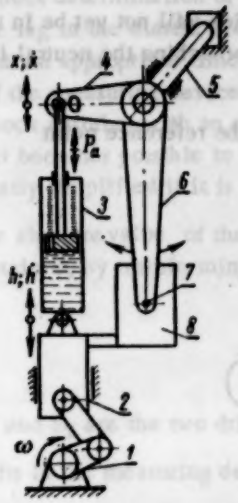

Fig. 1. Schematic of the shock absorber test rack. 1) Crankshaft-rocker mechanism; 2) slider; 3) shock absorber; 4) elastic rod lever; 5) torsion rod; 6) recording lever; 7) pencil (writing stylus); 8) board for recording the resisting effort to the piston movement.

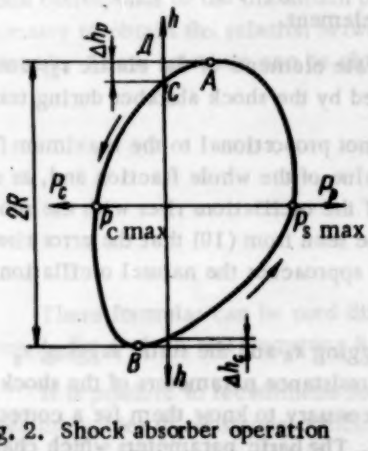

Fig. 2. Shock absorber operation diagram recorded on a dynamometer test rack. P_s and P_c are the resisting forces to stretching and compression; Δh is the lag between the oscillations of the elastic measuring device and the movement of the drive; ΔP is the deviation of the elastic measuring device from the neutral line due to the lag in its oscillations; 2R is the full displacement of the drive slide.

$$P_a = k \cdot v^i, \tag{1}$$

where P_a is the resisting force of the shock absorber; k is the coefficient of proportionality; v is the speed of vibrations; i is the exponent.

In order to analyze the measuring system of the dynamometer test rack, it is convenient to consider it as an oscillatory system excited by the resisting force of the shock absorber.

Then we can write on the basis of the Lagrange-d'Alembert theorem:

$$M\ddot{x} + cx = P_a, \tag{2}$$

where x is the elastic element displacement at the point where it is fixed to the shock absorber; M is the mass referred to this point; c is the stiffness referred to the same point; P_a is the shock absorbing force.

Here we neglect the insignificant friction which, however, always exists in the system, and we also assume that the elastic system does not have any liquid damping of its own. Let us find the value of the exciting force. If we adopt the generally accepted hypothesis that the relation between the shock absorber resisting force and speed is linear, we shall obtain from the general equation (1):

$$P_a = kv. \tag{3}$$

In the latter expression it is necessary to define more precisely the value of the relative velocity v with respect to the test conditions. It is clear from the principle of operation of the shock absorber (see Fig. 1) that both ends of the shock absorber are moving, the lower end at the velocity $\dot{h}$ of the vibrator slider, and the upper end at the velocity $\dot{x}$ of the measuring lever, and that both movements are in the same direction. The velocity of the relative movement conveyed to the shock absorber is, therefore,

$$v = \dot{h} - \dot{x}. \tag{4}$$

Let us replace in the latter expression the value of $\dot{h}$ by the parameters of the installation's mechanical drive

$$\dot{h} = R\omega \cdot \sin \omega t,$$

where R is the radius of the crank; ω is the angular frequency of oscillations; t is the time.

We assume for the sake of simplicity that the movement is harmonic. Let us now insert the value of P_a into (2):

$$M\ddot{x} + k\dot{x} + cx = kR\omega \sin \omega t. \tag{5}$$

Since in these tests it is required to determine the resisting force by producing forced oscillations in the elastic system, the general solution of (5) is of no interest to our investigation. Its particular solution for forced oscillations should be obtained according to the theory of oscillations [4] in the form:

$$x = x_0 \sin(\omega t - \varphi). \tag{6}$$

Angle φ is the measure of the phase displacement in the measuring device oscillations with respect to those of the drive, which occur at a frequency of ω. The phase shift angle φ is a negative quantity. This indicates that the elastic element's oscillations lag by angle φ from the drive oscillations. In other words, when the vibrator slide is, for instance, at the end of the shock absorber's expansion stroke (in the lower dead center) the slide's velocity is zero. Therefore, the shock

absorber's resisting force is also zero, but at that moment the elastic measuring device will not yet be in the neutral position and, due to its lag, will have a certain deviation, namely, it will only be approaching the neutral line (points A and B in Fig. 2).

Hence, by solving (5) we shall obtain the maximum oscillation amplitude at the reference point

$$x_0 = \frac{kR\omega}{\sqrt{(k\omega)^2 + (c - M\omega^2)^2}}, \tag{7}$$

or

$$x_0 = \frac{\frac{kR\omega}{c}}{\sqrt{(1-i^2)^2 - \left(2\frac{k}{k_a}\cdot i\right)^2}} = \frac{x_s}{\sqrt{(1-i^2)^2 - \left(2\frac{k}{k_a}\cdot i\right)}} \tag{8}$$

and the tangent of the phase shift angle:

$$\operatorname{tg}\varphi = \frac{k\omega}{c - M\omega^2}, \tag{9}$$

or

$$\operatorname{tg}\varphi = \frac{2\frac{k}{k_a}\cdot i}{1 - i^4}, \tag{10}$$

where k_a is the resistance coefficient which produces the aperiodic movement of the oscillatory (elastic) system $k_a = 2\sqrt{Mc} = 2M\omega_0$; ω_0 is the natural oscillation frequency of the system; x_s is the static deformation of the elastic element under the effect of force $kR\omega$, which is the maximum resisting force of the shock absorber under ideal conditions when there is no sagging in the elastic element, i.e., when its bracing is perfectly stiff; $i = \omega/\omega_0$ is the ratio of the excitation frequency to the natural frequency of the elastic element.

Let us now examine how the above expressions reflect the effect of separate elements of the elastic system and the resistance parameters on the accuracy of recording the efforts developed by the shock absorber during testing.

It will be seen from (7) that the maximum sag in the elastic element is not proportional to the maximum force $P_{a\,max}$, since the denominator contains variable quantities which affect the value of the whole fraction and, as experience shows, cannot be neglected. It follows from (9) that the phase shift of the oscillations rises with the shock absorber resistance coefficient k, i.e., the error increases with k. It will also be seen from (10) that the error rises with the frequency of driving oscillations, i.e., when the excitation frequency approaches the natural oscillation frequency of the measuring device.

From (8) it is possible to determine the relation between the dynamic sagging x_0 and the static sagging x_s which is produced by the static effect of the force. This relation includes the resistance parameters of the shock absorber and the parameters of the oscillatory measuring system; hence, it is necessary to know them for a correct evaluation of the results when testing shock absorbers with unknown parameters. The basic parameters which characterize the elastic measuring device consist of the stiffness of the measuring element and its mass or moment of inertia; hence, they must be determined first of all. The stiffness of the system can be obtained from a static calibration (or a theoretical computation of the elastic element), its mass can be determined theoretically or experimentally; in the latter case a special dynamic calibration is required with subsequent computations by means of equations (7-10).

Dynamic calibration amounts in essence to determining the natural oscillation frequency of the elastic measuring device. In practice this can be attained by direct or indirect measurements.

In the first instance measurements are made by means of a normal vibrograph, which is fixed* to lever 4 of the measuring device (see Fig. 1). The natural frequency of oscillations of the system is produced artificially by twisting the torsion bar and then suddenly releasing it.

* If the vibrograph is mounted in such a manner that the lever supports the whole of its mass, the latter should be subtracted for greater accuracy from the measured mass.

Indirect determination of the frequency is based on an analysis of the value of Δh (see Fig. 2), which characterizes the lag in the elastic measuring device oscillations with respect to the drive oscillations. The value of Δh provides, on an appropriate time scale, the value of the oscillations' phase shift, which is related to the natural frequency of the measuring device by equation (10). It follows from the above that by testing on the installation a special shock absorber with an equal absolute value of resistance in compression and expansion and a linear relation to speed, it becomes possible to evaluate the natural frequency of the measuring device in the installation. This testing is greatly simplified if it is possible to carry it out at two different frequencies of drive oscillations.

The absolute value of the shock absorber's resistance coefficient is then eliminated and the following relationship is obtained, by transforming equation (10), and used for calculating the natural frequency of oscillations:

$$\omega_0 = \sqrt[4]{\frac{\omega_1^4 - A\omega_2^4}{1 - A}}, \tag{11}$$

where ω_1 and ω_2 are the two drive oscillation frequencies used for testing; $A = \frac{\omega_1}{\omega_2}\frac{\operatorname{tg}\varphi_2}{\operatorname{tg}\varphi_1}$, $\operatorname{tg}\varphi_1$ and $\operatorname{tg}\varphi_2$ are the phase shifts in the measuring device oscillations corresponding to the above test frequencies.

The angle of the measuring device oscillations phase shift is

$$\varphi = \arccos\left(\frac{R - \Delta h}{R}\right). \tag{12}$$

Knowledge of the elastic measuring device oscillatory parameters (mass, elasticity, natural frequency and coefficient of aperiodicity k_a) makes it possible to evaluate in practice the errors arising in testing shock absorbers by means of this installation.

Normally both at home and abroad [5] shock absorbers are evaluated by their maximum resisting force at various testing conditions. It has already been shown that x_0 is the maximum deviation of the elastic measuring device, which corresponds to the maximum resisting force of the tested shock absorber. Hence, for practical purposes it is necessary to obtain the relation between the dynamic sagging and the actual value of force P_a or the resistance coefficient k. These relations can be obtained by transforming (7)

$$k = \frac{x_0(c - M\omega^2)}{\omega\sqrt{R^2 - x_0^2}}, \tag{13}$$

$$P_a = \frac{Rx_0(c - M\omega^2)}{\sqrt{R^2 - x_0^2}}. \tag{14}$$

These formulas can be used directly for evaluating as a first approximation the resistance coefficient k and force P_a for a given test frequency $R\omega$.

It is possible to recommend for day-to-day use both in laboratories and production a special table which provides the values of resistance coefficient k and force P_a for various values of a maximum sag, x_0.

By means of another transformation, (7), it is possible to obtain a formula for determining the actual maximum velocity of the shock absorber piston under test.

$$v_{max} = \frac{R\omega}{\sqrt{\left(\frac{k\omega}{c}\right)^2 + \left(\frac{M\omega^2}{c}\right)^2 - \frac{2M\omega^2}{c} + 1}} \tag{15}$$

So as to provide a better idea of the errors which arise in testing, the drop in the maximum velocity of the shock absorber piston with respect to the shock absorber resistance coefficient is given in Fig. 3. The installation has the following parameters: R = 3.8 cm, ω = 10.4 sec^{-1}, c = 180 kg-wt/cm, mass M is neglected owing to its small value.

On the basis of the above relationship it is possible to plot a nomogram for determining the actual value of the resisting force and the test velocity from the measurement results. The principle for plotting such a nomogram is shown in Fig. 4.

The resisting force is plotted on the vertical axis of the coordinate system, the velocity or frequency of oscillations which may occur during testing are plotted along the horizontal axis. From the origin of the coordinates straight lines are drawn which relate the values of P_a and $\underline{v}$ with the resistance coefficients k_1, k_2, k_3, etc., according to (3).

For each possible test velocity v_1, v_2, v_3. . . ., curves 1, 2, 3. . . . are plotted which are bent toward the P_a axis, thus indicating a drop in the actual velocity of the relative movement of the shock absorber piston with a rising resisting force.

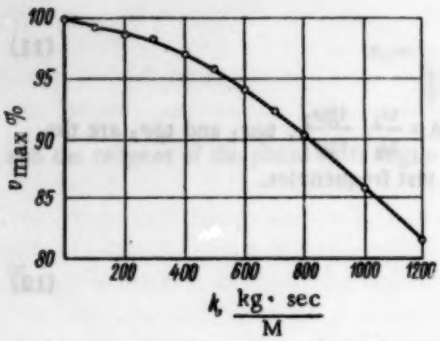

Fig. 3. Fall in the relative velocity of the shock absorber piston for various resistance coefficients used in testing on the dynamometer installation.

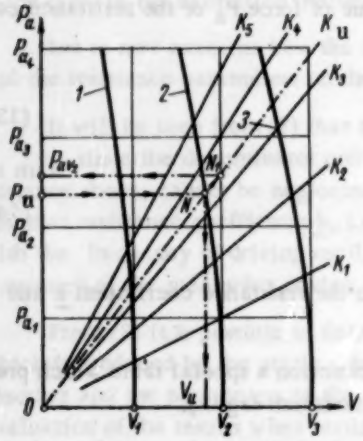

Fig. 4. Schematic of a nomogram for evaluating the results obtained in testing shock absorbers on a dynamometer installation.

These curves are plotted from an equation obtained by transforming (15)

$$v = \frac{R\omega}{\sqrt{\left(\frac{P_a}{Rc}\right)^2 + \left(\frac{M\omega^2}{c}\right) - \frac{2M\omega^2}{c} + 1}} \qquad (16)$$

The nomogram is used in the following manner: the resisting force is determined on the operation diagram from the static calibration data. The value of this force, for instance P_u, is then found on the nomogram and a horizontal line is drawn from it up to its intersection with the curve corresponding to the test velocity, for instance v_2. The point N thus obtained in the coordinate field makes it possible to find the actual test velocity v_u by dropping from point N a perpendicular onto the $\underline{v}$ axis, and also the value of the resisting force P_a for the case when the test velocity is actually equal to v_2.

For the latter purpose it is necessary to draw a straight line k_u from the origin of the coordinates through point N up to its intersection with the vertical line passing through v_2. A projection from the intersection point N_1 onto the P_a axis will provide the value of the resisting force corresponding to velocity v_2.

It should be noted that the above method provides accurate results only for a linear relationship between the resisting force and velocity of oscillations. The greater the deviations of index $\underline{i}$ in (1) from unity, the more approximate become the evaluations of parameter P_a and $\underline{v}$.

However, for the quadratic relationship ($i = 2$), most commonly encountered in practice, the above method is sufficiently accurate This is confirmed by comparing the hydraulic investigations data with test results obtained on a dynamometer test rack [6]. Hence, in using the nomogram when $i = 2$, it is necessary to remember that the value of P_u represents an instantaneous resisting force and is determined by

$$P_u \approx k_2 \cdot v_u^2. \qquad (17)$$

where $k_2 \approx k_u / v_u$ is the resistance coefficient for a square law.

The values of P_a and $\underline{v}$ in shock absorbers whose resisting force follows a square law can also be obtained by solving the differential equation (2), in which the excitation force should be expressed in the form

$$P_a \approx k_2(\dot{h} - \dot{x})^2. \qquad (18)$$

Such a solution, however, requires different methods from those previously used, and at the same time does not substantially raise the accuracy of determining P_a and $\underline{v}$ as compared with the above-mentioned method.

By means of a further analysis of the above equations it is possible to determine the limiting resisting force which can be recorded by the measuring element of the installation

$$P_{a\,\max} = Rc. \qquad (19)$$

In practice the force should be limited to

$$P_{a\ max} < 0.5\ Rc. \qquad (20)$$

On the basis of this value design computations can be made for the strength and fatigue of the elastic element in new dynamometer installations for testing shock absorbers.

The smallest resisting force which can be measured by means of this installation is determined from the shock absorber's minimum resistance coefficient k_{min} which must not be smaller than k_a, the resistance coefficient which provides an aperiodic damping of the elastic measuring device natural oscillations. Otherwise the natural oscillations of the measuring device will distort the recording of the operation diagram.

It can also be shown that the greater the stroke and the smaller the frequency of oscillations for the same maximum velocity, the more accurately will the operation diagram be plotted. This accuracy is also increased by raising the stiffness of the elastic element. In the latter case, however, the recording scale must be suitable for the analysis of the diagram. In order to meet this requirement it is necessary to choose the ratio between the thickness of the line drawn on the board and the deviation of the pencil from the neutral line not less than $^1/_8$.

The above investigations have shown that in testing oscillation dampers of dynamometer installations, errors will always arise due to the inherent defect in the elastic oscillatory system which is used as a dynamometer. This defect consists of the discrepancy between the measurement results obtained with static and dynamic forces. The error due to this discrepancy rises with the vibration velocity of the force.

Moreover, the elastic measuring system is affected to a great extent by friction, i.e., by processes involving the absorption or dissipation of energy. Thus, the application of an elastic system for measuring shock-absorbing devices is peculiar in the way that the measured variable has a detrimental effect on the measuring device. In this connection it appears that the greater the resistance of the shock absorber, the larger becomes the error involved in its measurement.

Attempts to avoid these errors lead to conflicting effects. For instance, the raising of the elastic element stiffness requires an increase in the recording gear ratio, which in turn increases the effect of inertia and friction on the sensing elastic element in the recording mechanism.

Moreover, the intermediate transmitting elements of the system may, by virtue of their own stiffness, cause additional errors, particularly evident in raising the phase shift angle of the oscillations. This effect is sometimes incorrectly interpreted as being due to a special elastic property of the shock absorber.

When the same oscillation damper is tested on different installations whose elastic elements differ in their oscillatory parameters but whose remaining conditions are the same, it is possible to obtain discrepancies in the measurement results owing to the above-mentioned peculiarities. Such discrepancies occur in practice.

The above reasoning leads to the conclusion that it is necessary to provide special dynamometer installations with a measuring system which could serve for reference purposes.

At present the conditions are ripe for standardizing dynamometer methods of testing oscillation dampers. It is in the interests of our industry to develop uniform technical requirements for the measuring equipment intended for this purpose.

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INSTRUMENT FOR EXPERIMENTAL VERIFICATION OF THE HYDRODYNAMIC THEORY OF LUBRICATION IN ROLLING

A. A. Kut'kov

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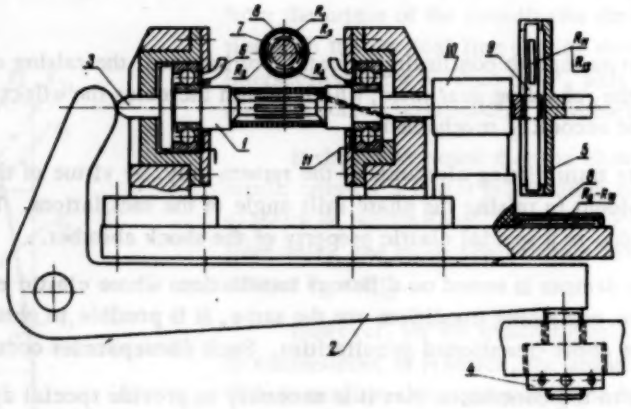
For the experimental verification of the hydrodynamic theory of lubrication in rolling, the author of this article has developed an instrument which provides direct measurements of the oil film thickness in working rolling bearings as a function of the viscosity of the lubricant, the load, the speed, etc. The basic premise for developing the instrument consisted in the idea that between the rolling components and the race a hydrodynamic flow is generated which separates the components from the race. Hence, if the external race of the bearing is rigidly fixed and the bearing is loaded with a radial load, the internal race must at certain speeds change its original position under the effect of the lubricant flow.

Considering that the setting of races would complicate the deciphering of the experimental results, in the proposed instrument the effect of this factor is eliminated by using radial-thrust bearings with an axial load. Thus, if two radial-thrust bearings are mounted on a shaft (see figure), their external races are rigidly fixed preventing any axial displacement, and a lubricant is supplied to the moving shaft, the lubricant flow thus formed will deform the



shaft producing in it a compression stress. By measuring the value of the deformation it is possible to determine the thickness and strength of the oil film.

The mechanical part of the instrument consists of a steel shaft 1 mounted in the radial-thrust ball bearings under test. The external races are mounted in barrels with a close running fit, which makes it possible for them to be displaced in the axial direction up to the stops. The barrels are fixed in brackets, which in turn are fixed to the table by means of bolts. Shaft 1 is loaded through lever 2, stop 3 and the left-hand side bearing by means screw mechanism 4. The screw mechanism has a fine thread which, in conjunction with the lever system, provides a smooth loading of the shaft. When the nut is rotated through $^1/_{50}$ of a turn (as read on a dial), stop 3 is displaced in an axial direction by 1 μ and provides a 10 kg-wt load on the bearing. All the components of the instrument subject to deformation as a result of the shaft loading have been provided with such dimensions, materials and heat-treatment as to make it possible to consider them completely rigid with respect to the shaft. The shaft is driven by an electric motor through a flexible coupling 5. The lubricant is supplied to the bearings through pipes 6 which are fixed to the lids, and the bearings are drained through holes which are closed by means of stoppers placed in the lower part of the lids.

The electrical part of the instrument comprises two bridges consisting of resistance transducers fed with an audiofrequency current through voltage stabilizers. The bridge readings are displayed on a loop oscilloscope MPO-2, which uses a current amplifier.

In order to measure shaft compression deformations and at the same time compensate for temperature deformations which may arise in the body of the shaft during the experiment, the author of this article has developed a bridge circuit consisting of eight identical transducers, with two transducers connected in series in each arm. The operating transducers R_1 and R_4 are glued on the shaft along its axis in order to measure its normal tension only. The compensation transducers R_2 and R_3 are glued to thin steel plates 7, which are placed over the whole length of the shaft, and transducers R_5, R_6, R_7, R_8, R_9, and R_{10} are mounted on an unloaded part of the steel table. In order to eliminate the effect of random external heat flows the measuring bridge transducers R_1, R_2, R_3, R_4, R_5, R_6, R_7, R_8, R_9, and R_{10} are lagged by means of thermal insulators 8. Tests of the instrument's measuring system have shown that the bridge provides stable readings, i.e., no zero drift was observed during the whole of the testing (15 min) of the instrument under load and without lubrication. Since the time taken by the testing does not exceed ten seconds at intervals of one hour, it can be assumed that the measuring system operates with the required accuracy.

Transducers R_{11} and R_{12} are glued to an elastic spring 9 of coupling 5 for the purpose of determining the moment of friction of the bearings before, during, and after lubrication. The supply and measuring leads are taken through a mercury connector 10. The temperature of the lubricant is measured by means of two thermometers 11. The rotation speed of the shaft is controlled by a smooth variation of the motor current, and measured by means of a stroboscopic method.

Tests have shown that the instrument is completely suitable for thorough experimental checking of the hydrodynamic theory of lubrication in rolling bearings.

MEASUREMENTS OF TIME

TRANSISTORIZED PULSE AMPLIFIER FOR A PRINTING CHRONOGRAPH

E. I. Dolbak

Translated from Izmeritel'naya Tekhnika, No. 8,
pp. 19-20, August, 1961

The printing chronographs type 21P made by the ÉChL plant are used by the time services of the Soviet Union in astronomical determinations of time corrections by means of transit instruments. The chronograph is designed in such a manner that its operation and printing on the tape is controlled by mechanical breaking and making of the input registering circuit of the chronograph. The instruments input current amounts to 25-30 ma.

In determining corrections for clocks it is necessary to record on the chronograph tape both the instant when the micrometer contacts of the transit instrument break and the instant when the electric minute pulse starts in the clock whose correction is being determined. When the micrometer contacts are connected to the input of the chronograph they burn out rapidly, due to the large chronograph input current. The conversion of electrical pulses of the clock into mechanical making and breaking by means of sensitive polarized relays results in additional errors when recording minutes and is unstable in operation owing to the large input current of the instrument.

The attached figure shows the schematic of a pulse amplifier which uses junction transistors and is made in the form of an attachment fixed to the chassis of the printing chronograph. The attachment provides a recording of the mechanical breaking of the transit instrument micrometer contacts and of positive electrical pulses of an amplitude greater than 0.5 v.

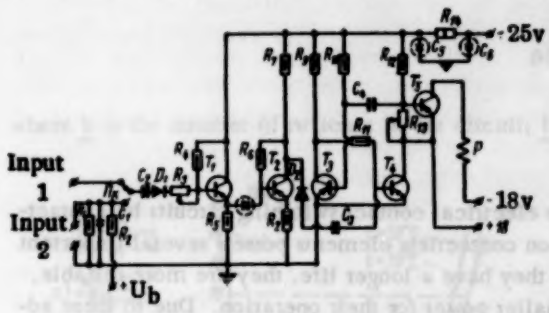
The contacts of the transit instrument micrometer are connected to input 2 of the circuit shown above. The mechanical breaking of the micrometer contacts is converted by means of the potential divider consisting of resistors R_1 and R_2 into positive voltage pulses which are fed through switch S to the amplifier input. With a suitable choice of the values of resistors R_1 and R_2 and voltage U_b, the current through the micrometer contacts can be made sufficiently small. The positive pulses from the clock are fed to input 1 and transmitted through switch S to the amplifier input.

The first stage has a large input impedance and consists of a grounded collector transistor T_1. The variable resistor R_3 in the base of this transistor serves to vary the sensitivity of the amplifier. The transistor T_2 amplifying stage uses both a dc and ac negative feedback. The amplified negative pulses are fed from collector T_2 through diode D_2 and capacitor C_3 to the input of the trigger with a single stable condition consisting of transistors T_3 and T_4. The controlled trigger is required for standardizing the duration and amplitude of the input pulses. The base-emitter junction of the keying transistor T_5 is connected to the collector circuit of trigger T_4. The former transistor operates in the "on-off" condition. Its load consists of a high-speed relay P, whose contacts supply the voltage to the windings of the chart-propelling and the printing mechanism of the chronograph.

In the initial condition T_3 is conducting and T_4 is blocked. The base current of T_5 which forms part of the collector current T_4 is small, and the collector current is lower than the relay operating current. An incoming pulse blocks T_3 and makes T_4 conducting for a time mainly determined by the time constant of $R_{10}C_4$. The base current of the keying transistor rises and its collector current reaches a value sufficiently large for a reliable operation of the relay.

The variable resistance R_{12} serves to control the value of the base current and, hence, the collector current of T_5. In order to make this current as small as possible in the "blocked" condition it is necessary to select transistors for the trigger which have a small initial collector current I_{co}. Moreover, it is necessary to select a trigger operating condition for which the collector current of T_4 should be as small as possible in the blocked condition. Depending on the selected circuit parameters, the collector current of T_5 changes from 5 to 35 ma for an operating current of relay P varying between 8-15 ma.

The trigger pulse duration required for a reliable operation of the chronograph chart-propelling and printing mechanism depends on the adjustment of relay P and lies in the range of 50-100 msec.



The maximum gain of the amplifier is determined by the ratio between the input signal power and the maximum power produced in the winding of relay P, and is of the order of 10^6.

T_5 is fed from a separate rectifier, in order to avoid dc coupling between the transistor load circuit and the remaining portion of the amplifier. The rectifier step-down transformer uses a type Sh-20 iron core 30 mm thick. The rectifiers are of the bridge type and use DG-Ts24 diodes.

The transistorized pulse amplifier is also suitable for recording on the printing chronograph the beating of two compared frequencies. The beat frequency voltage is fed to input 1 of the amplifier. From the distance between two adjacent imprints on the chronograph chart it is possible to determine the frequency of the beats with an accuracy up to 5 msec. The amplifier can also be used under other conditions where it is required to convert weak current pulses into powerful ones with a small delay time.

I INTERNATIONAL CONGRESS OF CHEMICAL ENGINEERING, TECHNOLOGY, AND AUTOMATION

On September 3 to 8, 1962, the I International Congress on Chemical Engineering, Technology and Automation will be held in Brno (Czechoslovakia).

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All the papers will be translated simultaneously into Russian, English and German.

The organizational committee of the I International Congress on Chemical Engineering, Technology and Automation is located at the exhibition pavilion No. 1 in Brno.

ELECTRICAL MEASUREMENTS

DIODE SWITCHING CIRCUITS FOR MEASURING PURPOSES

R. R. Kharchenko and V. N. Malinovskii

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In automatic control technique there is a tendency to replace electrical contact switching circuits by contactless circuits. This is due to the fact that controlled switches based on contactless elements possess several important advantages as compared with electromechanical switches; namely, they have a longer life, they are more reliable, they have a higher operating speed, a smaller size and require a smaller power for their operation. Due to these advantages contactless switches are beginning to be used successfully in many automatic devices, for instance, in telemetering circuits for modulating the transmitted signal, in automatic digital bridges and potentiometers for a stepped variation of the balancing resistances, in instruments with centralized test scanning for switching transducer circuits [1, 2, 3].

In the present work we examine the methods of computing switching circuits which use semiconductor diodes* and are designed for contactless connection and disconnection of resistors in a measuring circuits containing a direct current or voltage source. The main object of this work consists in determining the statistical errors due to switching, which arise in the measuring circuit parameters established by means of switching.

If the dynamic properties of the diode switching circuits are not important, the best element for its construction consists of a silicon junction diode; therefore, such elements were used in the experimental part of this work. The general approach to a comparative analysis of switching circuits (static) is applicable, however, to circuits with other types of diode rectifiers.

Figure 1 shows a typical volt-ampere characteristic of a silicon junction diode D-202. It is known that current $\underline{i}$ through a blocked diode depends but little on the value of the reversed voltage when the latter varies over wide limits, but it is very sensitive to changes in the temperature of the diode. Therefore, it is convenient to consider a blocked diode as a generator of a reversed current $\underline{i}$ whose value depends only on temperature (Fig. 1b). It is convenient to consider the diode in the conducting direction as an equivalent emf $\underline{e}$ and the diode's dynamic resistance $\underline{r}$ connected in series (Fig. 1a). The values of $\underline{e}$ and $\underline{r}$ depend on the position of the operating point on the volt-ampere characteristic, and are determined graphically as shown in Fig. 1. It will be seen from the volt-ampere characteristic shown in Fig. 1 that the variations of temperature produce insignificant changes in $\underline{r}$, but emf $\underline{e}$, on the contrary, varies considerably and has a negative temperature coefficient. Figure 2 shows the relationships $r = f(I)$ and $e = \varphi(I)$ for the silicon diode D-202 at a constant temperature. It will be seen from the graphs that for a sufficiently large current I through a conducting diode, small variations of this current produce small changes in parameters $\underline{r}$ and $\underline{e}$. Hence, for such a condition these parameters can be considered approximately constant.

Let us first examine a circuit for switching resistances in a supply circuit. Switching circuits of this type are used in digital to analogue converters based on the summation of stable currents [2]. The most typical version of such a converter is illustrated by the circuit in Fig. 3. The circuit consists of resistors connected in series according to the law R, R, 2R, 4R, . . . 2^{k-2} R. The nodal points, $\underline{a}$, $\underline{b}$, $\underline{c}$, etc. of this circuit are connected through contactless switches D_1-D_2, D_3-D_4, . . . D_{2k-1}-D_{2k} to current sources I, for instance, cascade triode circuits operating as current stabilizers. The switches are controlled by triggers T which use electron tubes. The circuit has two additional supply sources with voltages E_c and E, with the controlling voltage E_c being larger than voltage E. In the initial condition all the triggers T are in the 0 position. The current from source E_c flows through diodes with even subscripts, namely D_2, D_4, . . . D_{2k}, and the current generator I. The potentials at points $\underline{e}$, f, g, . . . $\underline{n}$ of the

*There also exist circuits which use transistors.

**The above problem, to the best of the authors' knowledge, is being dealt with in print for the first time.

circuit thus become higher than those at points a, b, c, . . . m, respectively. Under this condition the diodes with odd subscripts, namely, D_1, D_3, . . . D_{2k-1}, are blocked, which makes the switches nonconducting. The reversed currents of the latter diodes provide at the output of the circuit a voltage of

$$\delta U = iR \sum_{p=1}^{\kappa} 2^{p-1},$$

where k is the number of switches in the circuit; i is the current through a blocked diode.

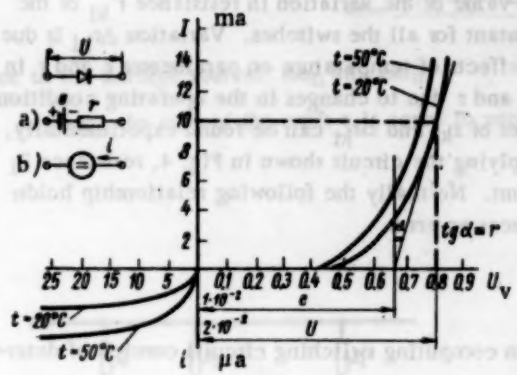

Fig. 1.

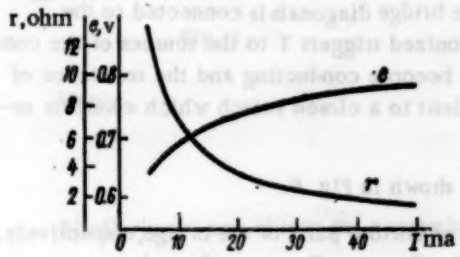

Fig. 2.

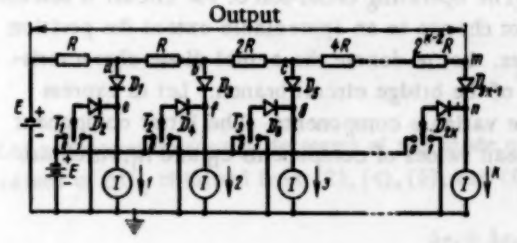

Fig. 3.

If trigger T_1 is now switched to position 1 and the rest of the triggers remain in position 0, diode D_2 will become connected (in the nonconducting direction) in parallel with current generator I and current I' = I + i will flow through diode D_1. Voltage ΔU = I'R will then appear across resistor R, but the voltage across the circuit output will differ slightly from ΔU:

$$\Delta U = \Delta U - (\delta U - iR) = iR\left[1 - \frac{i}{I}\left(\sum_{p=1}^{\kappa} 2^{p-1} - 2\right)\right].$$

It is obvious that the second term in the square brackets represents the relative error introduced by the blocked switches into the value of the output voltage when the first switch is conducting. The above error refers to the worst possible condition, when only the first switch is conducting and the output voltage is equal to the minimum discrete unit of the converted scale.

In practice it is usually assumed that the permissible error of a digital to analogue converted should be made smaller or equal to half the minimum discrete unit of the scale. Let us adopt this condition, then we shall find the value of current I from the relation:

$$\frac{i}{I}\left(\sum_{p=1}^{\kappa} 2^{p-1} - 2\right) < 0.5,$$

where

$$I > 2i \sum_{p=2}^{\kappa} 2^{p-1}.$$

Let us assume for an example that the number of switches (and hence the number of binary places) is k = 10, and the reverse current of a silicon diode is $i \leq 5 \cdot 10^{-8}$ amp. With these assumptions I ≥ 100 μa.

In the case when all the triggers T of the circuit shown in Fig. 3 are in position 1, the diodes with even subscripts D_2, D_4... . . . D_{2k} become connected in parallel with current generator I. All the switches are blocked and currents I' = I + i flow through the diodes with odd subscripts D_1, D_3, . . . D_{2k-1}. The output voltage of the circuit becomes:

$$U = I'R \sum_{p=1}^{\kappa} 2^{p-1} = iR \sum_{p=1}^{\kappa} 2^{p-1}\left(1 + \frac{i}{I}\right).$$

If in this example the reversed diode currents do not exceed $5 \cdot 10^{-8}$ amp, and it is required to keep the error due to blocked switches below half the minimum discrete unit of the scale, it is obviously necessary to satisfy the relation I ≥ 100 μa.

The above circuit is characterized by the fact that none of its contactless switches are connected in series with the switched measuring circuit resistances. Therefore, the equivalent parameters of the switching diodes e and r do not affect the digital to analogue converter accuracy, which is determined exclusively by the degree of stabilization of currents $I_1, I_2, \ldots I_k$. All the other circuits examined below do not possess this property. In them the switches are connected in series with the switched measuring circuit resistances, and therefore the equivalent diode parameters e and r, which control the contactless switch resistances, affect the digital to analogue converted output signal.

Let it be required to set up resistances between points a and b by means of contactless switches in the circuit shown in Fig. 4. It is obvious that with a conducting switch the actual resistance in the ith branch of the circuit will be $R'_i = R_i + r_{ki}$, where r_{ki} is the equivalent resistance of the switch. Let us assume that r'_{ki} consists of two components, namely $r'_{ki} = r_{ki} + \Delta r_{ki}$, where Δr_{ki} is the maximum value of the variation in resistance r'_{ki} of the "blocked" switch as compared with a certain value r_{ki} which is constant for all the switches. Variation Δr_{ki} is due to the differences in the volt-ampere characteristics of diodes, the effects of temperature on parameters e and r in the equivalent circuits of various diodes, variations in parameters e and r due to changes in the operating conditions of diodes caused by variations in the controlling voltage. The values of r_{ki} and Δr_{ki} can be found experimentally, or calculated approximately by the method suggested below. In applying the circuit shown in Fig. 4, resistance R_i must be chosen with the constant component of r_{ki} taken into account. Normally the following relationship holds: $r_{ki} \ll R_i$. Nevertheless, the switch resistance variation Δr_{ki} introduces an error

$$\lambda = \frac{\Delta r_{ki}}{R'_i}.$$

From the above definitions it follows that the main problem in computing switching circuits consists of determining the values of r_k and Δr_k.

Let us examine two types of contactless semiconductor diode switches used for switching resistance R (subscript i is omitted here and will be henceforth omitted) of the circuit shown in Fig. 4.

The switch of the first type [3] is shown in Fig. 5. One of the diode bridge diagonals is connected to the switched resistance R, and the other through two resistances R_0 and synchronized triggers T to the sources of the controlling voltage E_c. When the triggers T are set in position 1, the diodes become conducting and the resistance of the bridge circuit between terminals b and c becomes very small, equivalent to a closed switch which connects resistance R to the circuit of the voltage source U.

In this condition Fig. 5 can be represented by its equivalent circuit shown in Fig. 6.

Controlling currents $(I_c)_t$ and $(I_c)_b$, which flow through to the top and bottom parts of the bridge, respectively, and whose sum can be represented by $I_c = E_c/R_0$ if the parameters of diodes $D_1 \ldots D_4$ are neglected.

Currents I_l and I_r flow through the right- and left-hand sides of the bridge, respectively, and can be represented by $I = U/R$, if the parameters of diodes $D_1 \ldots D_4$ are neglected. The operating condition of the circuit is selected so that relation $I \ll I_c$ holds, i.e., the existence of current I does not change to an appreciable extent the position of the operating point on the volt-ampere characteristic of the diodes. Let us denote the actual diode characteristic parameters by e'_n and r'_n, where subscript n denotes the number of the bridge circuit branch. Let us express these parameters by the sum of two components, the constant and the variable components. The latter component must represent the deviation of the values of e'_n and r'_n from the mean values of components e_n and r_n, which are selected to be the same for all four diodes of the circuit.

Then $e'_1 = e + \Delta e_1$; $e'_2 = e + \Delta e_2$ etc.,

Similarly $r'_1 = r + \Delta r_1$; $r'_2 = r + \Delta r_2$ etc. (1)

All the quantities in these expressions can easily be found experimentally from the volt-ampere characteristic for a given batch of diodes. Moreover, the values of Δe_n and Δr_n must account for the dispersion of parameters in the series, their deviations due to changes in the controlling currents caused by feeding voltage variations, and their deviations due to temperature effects on the diode characteristics.

By using the above definitions we can substitute the equivalent circuit shown in Fig. 6 by that shown in Fig. 7. In this circuit emfs e, Δe_n and ΔU_n are connected in series with resistance r in each branch of the bridge. The sum of the first two emfs is equal to the value of e'_n in the circuit of Fig. 6 for each branch respectively; components

ΔU_n are expressed on the basis of the compensation principle by the product of the actual currents and the resistance Δr_n for each branch of the bridge. Thus:

$$\Delta U_1 = \Delta r_1 [(I_c)_t - I_1].$$

The second term in the square brackets is smaller by one order than the first and can, therefore, be neglected. The first term can be found directly from the circuit shown in Fig. 7 by means of the current superposition principle. According to this principle the control current in each branch of the bridge has two components, namely, components $I_c/2$ due to the voltage sources E_c, and component

$$\Delta I = \frac{\Delta e_1 + \Delta e_2 - \Delta e_3 - \Delta e_4 + \Delta U_1 + \Delta U_2 - \Delta U_3 - \Delta U_4}{4r}, \quad (2)$$

due to the voltage sources Δe_n and ΔU_n.

Since the sum of the emfs e is equal to zero it does not provide the value of the current in circuit a, b, c, d.

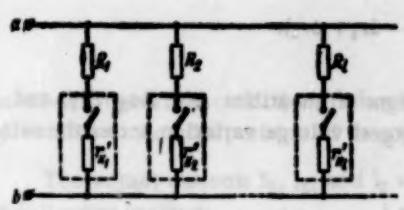

Fig. 4.

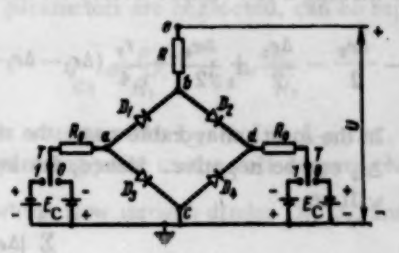

Fig. 5.

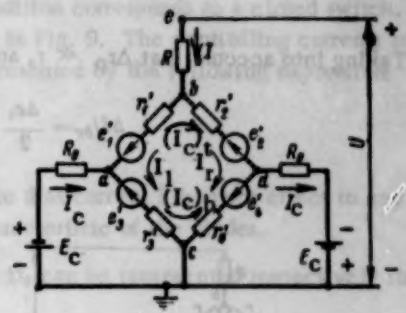

Fig. 6.

Hence,

$$\Delta U_1 \approx \Delta r_1 \left(\frac{I_c}{2} - \Delta I \right). \quad (3)$$

Similarly we can obtain the expressions

$$\Delta U_2 \approx \Delta r_2 \left(\frac{I_c}{2} - \Delta I \right), \quad (4)$$

$$\Delta U_3 \approx \Delta r_3 \left(\frac{I_c}{2} + \Delta I \right), \quad (5)$$

$$\Delta U_4 \approx \Delta r_4 \left(\frac{I_c}{2} + \Delta I \right). \quad (6)$$

Let us express current ΔI in terms of the diode parameter variations Δe_n and Δr_n. For this purpose let us insert the values of ΔU_n obtained from (3), (4), (5), and (6) into (2). We shall then obtain after simple transformations

$$\Delta I = \frac{\Delta e_1 + \Delta e_2 - \Delta e_3 - \Delta e_4 + \frac{I_c}{2}(\Delta r_1 + \Delta r_2 - \Delta r_3 - \Delta r_4)}{4r + \Delta r_1 + \Delta r_2 + \Delta r_3 + \Delta r_4}. \quad (7)$$

Now it is easy to obtain from the circuit of Fig. 7 the required constant component r_k of the switching circuit equivalent resistance and its variation Δ_{rk}.

We can find r_k by assuming all the emfs of the type of Δe_n and ΔU_n to be equal to zero. Then we shall see that $\Delta I = 0$ and the voltage between points b and c of the circuit is represented by expression

$$U_{bc} = \frac{1}{2} r (I_c + I) - \frac{1}{2} r (I_c - I) = Ir.$$

From the above it follows that:

$$r_к = \frac{U_{bc}}{I} = r. \qquad (8)$$

We can find the value of Δr_k from expression $\Delta U_{bc}/I$, where ΔU_{bc} is the voltage component between points b and c of the circuit shown in Fig. 7 due to the effect of Δe_n and ΔU_n type emfs only. By following round the circuit in Fig. 7 we obtain:

$$\Delta U_{bc} = \Delta e_1 - \Delta e_3 + \Delta U_1 - \Delta U_3 - \Delta I 2r.$$

Let us expand this expression by means of (3), (5), and (7)

$$\Delta U_{bc} = \Delta e_1 - \Delta e_3 + \frac{I_c}{2}\Delta r_1 - \frac{I_c}{2}\Delta r_3 -$$
$$- \frac{\Delta e_1 + \Delta e_2 - \Delta e_3 - \Delta e_4}{4r + \Delta r_1 + \Delta r_2 + \Delta r_3 + \Delta r_4}(2r + \Delta r_1 + \Delta r_3) -$$
$$- \frac{I_c}{2} \cdot \frac{\Delta r_1 + \Delta r_2 - \Delta r_3 - \Delta r_4}{4r + \Delta r_1 + \Delta r_2 + \Delta r_3 + \Delta r_4}(2r + \Delta r_1 + \Delta r_3).$$

Taking into account that $\Delta r_n \ll r$, and making a few simple transformations we obtain:

$$\Delta U_{bc} = \frac{\Delta e_1}{2} - \frac{\Delta e_2}{2} - \frac{\Delta e_3}{2} + \frac{\Delta e_4}{2} + \frac{r_y}{4}(\Delta r_1 - \Delta r_2 - \Delta r_3 + \Delta r_4).$$

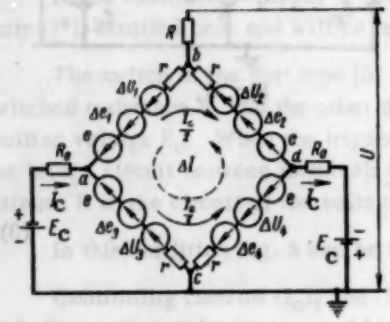

Fig. 7.

In the most unfavorable case the signs of quantities Δe_2, Δe_3, Δr_2, and Δr_3 can be negative. Hence, the largest voltage variation across the switch will be:

$$|\Delta U_{bc}| = \frac{\sum_{n=1}^{4} |\Delta e_n|}{2} + \frac{I_c}{4}\sum_{n=1}^{4} |\Delta r_n|.$$

Hence, the required maximum variation of the switch resistance will assume the form:

$$|\Delta r_к| = \frac{|\Delta U_{bc}|}{I} = \frac{1}{2I}\sum_{n=1}^{4}|\Delta e_n| + \frac{1}{4}\cdot\frac{I_c}{I}\sum_{n=1}^{4}|\Delta r_n|. \qquad (9)$$

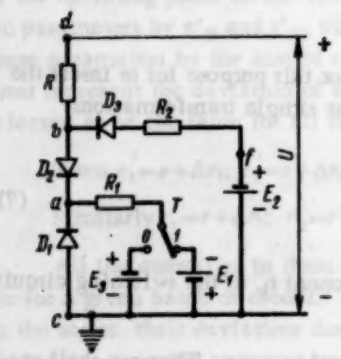

Fig. 8.

The value of $|\Delta r_k|$ may become larger than the constant equivalent resistance r_k of the switch as calculated from (2). This is confirmed by experimental data. The following figures provide an approximate idea of the parameter values for the switching circuit shown in Fig. 5 and consisting of silicon diodes D-202: I_c = 100 ma; I = 5 ma; e = 0.8 v; $(\Delta e_n)_{max}$ = = 0.05 v; r = 1.2 ohm; $(\Delta r_n)_{max}$ = 0.2 ohm. The two last parameters have been determined for the following conditions: a temperature range of 15-30°C, and control voltage variations of ±10%. The values of the required resistances as calculated from (8) and (9) are equal, respectively to r_k = = 1.2 ohm, and Δr_k = 22 ohm.

Let us now refer back to Fig. 5 for an analysis of a closed switch operating condition when triggers T are in position 0. In this condition all the diodes are blocked and reverse currents i_n flow through them. Hence, the circuit of resistance R has a current flowing through it which is equal to the difference between the reverse currents in D_2 and D_1, namely $i_R = i_2 - i_1$. The reverse currents of diodes D_3 and D_4 will not flow through the circuit of resistance R. Currents i_1 and i_2 in silicon diodes do not exceed the value of $5 \cdot 10^{-8}$ amp; hence, the current i_R will not exceed that value. Thus,

the equivalent resistance of an open switch can be represented by the expression

$$R_k = \frac{U}{i_2 - i_1}, \tag{10}$$

where U is the external voltage in the circuit of Fig. 8.

If its value is not less than several volts (digital bridges), then R_k will equal several hundred megohms, which in practice is of the order of the insulation resistance; moreover, the sign of R_k can be either positive or negative, depending on the sign of the difference $i_2 - i_1$.

In conclusion let us examine yet another type of a diode (differential) switch [1] whose schematic* is shown in Fig. 8. The switch consists of three diodes D_1, D_2, and D_3, two resistances R_1 and R_2, a source of auxiliary direct voltage E_2, and sources of control voltages E_1 and E_3 which are switched by trigger T.

The voltages of sources E_2 and E_3 are chosen to make the potential at point f smaller than that at point d, and the potential at point 0 (trigger T) higher than that at point d.

Let us assume that trigger T is in position 1. Then all the diodes of the circuit will be conducting and will pass a current of I_{C1} and I_{C2} due to E_1 and E_2 and current I due to U. This condition corresponds to a closed switch. Let us use the preceding method of analysis and represent the switch as shown in Fig. 9. The controlling currents in the first approximation, when the diode parameters are neglected, can be represented by the following expression:

$$I_{C1} = \frac{E_1}{R_1}; \quad I_{C3} = \frac{E_3}{R_3}.$$

The circuit operating condition is chosen so as to make $I_{C1} \gg I \ll I_{C2}$, i.e., so that current I does not affect to any appreciable extent the position of the operating point on the volt-ampere characteristic of the diodes.

The actual currents I_1, I_2, and I_3 which flow through diodes D_1, D_2, and D_3 can be represented respectively in the following manner:

$$I_1 = I_{C1} - I_{C2} - I;$$

$$I_2 = I_{C2} + I;$$

$$I_3 = I_{C3}.$$

Let us denote the actual diode characteristic parameters by e'_n and r'_n where the subscript n represents the number of the diode in the switching circuit. The values of e'_n and r'_n are expressed as before by the sums of $e + \Delta e_n$ and $r + \Delta r_n$, which have already been determined in (1) for the bridge switching circuit. Let us represent the circuit of Fig. 9 by its equivalent as shown in Fig. 10. Here the emfs ΔU_n are determined by means of the compensation principle

$$\Delta U_1 = \Delta r_1 I_1 = \Delta r_1 (I_{C1} - I_{C2} - I), \tag{11}$$

$$\Delta U_2 = \Delta r_2 I_2 = \Delta r_2 (I_{C2} + I) \tag{12}$$

$$\Delta U_3 = \Delta r_3 I_3 = \Delta r_3 I_{C3}. \tag{13}$$

The additional currents ΔI_n due to the internal sources Δe_n and ΔU_n can be neglected, since they are small compared with currents I_1, I_2, and I_3. We can find the resistance r_k of a closed switch by assuming all the emfs of the type of Δe_n and ΔU_n to be equal to zero. The voltage between points b and c can then be expressed as

$$U_{bc} = e + r(I_{C2} + I) - e - r(I_{C1} - I_{C2} - I) = r(2I_{C2} - I_{C1} + 2I) \tag{14}$$

If we assume that:

$$I_{C2} = \frac{1}{2} I_{C1},$$

*The circuit for this switch was suggested by V. N. Malinovskii.

then

$$U_{bc} = 2rI,$$

whence

$$r_к = \frac{U_{bc}}{I} = 2r. \tag{16}$$

Expression (15) can easily be met by an appropriate selection of parameters R_1, R_2, E_1 and E_2 in the circuit of Fig. 8. But the above circuit possesses an interesting peculiarity. By means of a simple adjustment of resistance R_1 or resistance R_2 it is possible to slightly change the relationship (15) and make the equivalent switch resistance r_k equal to zero.

In fact (14) and (16) indicate that r_k becomes equal to zero, providing

$$2I_{C_2} - I_{C_1} + 2I = 0$$

or

$$I_{C_2} = \frac{1}{2} I_{C_1} - I. \tag{17}$$

However, this stipulation is not met in all the operating conditions. Thus, if the circuit of Fig. 4 consists of an arm of a digital bridge, the voltage across a and b and, hence, the current I will depend on the number of closed switches. It is obvious that relation (17) cannot hold for an arbitrary number of connected resistances. The value of Δr_k can be found as before from expression $\Delta U_{bc}/I$, where ΔU_{bc} is the component of the voltage between points b and c of Fig. 10 due to voltages of the type of Δe_n and ΔU_n only.

These emfs in branches ab and ac of the circuit provide additional current components ΔI_1 and ΔI_2, which can be neglected since they are 1-2 orders lower than the controlling currents I_{C1} and I_{C2}. It can easily be seen that ΔU_{bc} will then take the form of the equation

$$\Delta U_{bc} = \Delta e_3 - \Delta e_1 + I(\Delta r_1 + \Delta r_3) + I_{C_2}(\Delta r_1 + \Delta r_3) - I_{C_1}\Delta r_1. \tag{18}$$

The third term in the right-hand side of this equation can be neglected since the circuit always operates under a condition when $I \ll I_{C_2}$. Moreover, it has already been noted that the approximate relation $I_{C1} = 2I_{C2}$ holds. Equation (18) can then be simplified to

$$\Delta U_{bc} = \Delta e_3 - \Delta e_1 + I_{C_2}(\Delta r_3 - \Delta r_1). \tag{19}$$

Considering that Δe_n and Δr_n can vary, we can write for the worst case

$$\Delta U_{bc} = \sum_{n=1}^{2} |\Delta e_n| + I_{C_2} \sum_{n=1}^{2} |\Delta r_n|.$$

Thus, the required expression for the maximum switch resistance variation Δr_k will finally take the form

$$|\Delta r_к| = \frac{1}{I} \sum_{n=1}^{2} |\Delta e_n| + \frac{I_{C_2}}{I} \sum_{n=1}^{2} |\Delta r_n|. \tag{20}$$

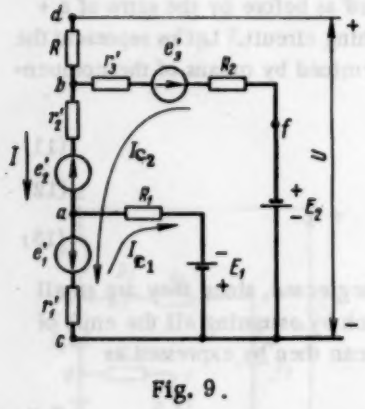

Fig. 9.

Let us now deal with the circuit of Fig. 8 and examine the operating condition of an open switch when trigger T is in position 0. All the diodes will then be blocked and reversed currents will flow through them.

The reverse current of diode D_1 does not flow through the switched resistance R, since its circuit is completed through source E_3. The difference of the reversed currents of diodes D_2 and D_3 flows through resistance R.

Hence, the equivalent resistance of an open switch in the circuit of Fig. 8 is represented by the expression $R_k = \frac{U}{I_3 - I_2}$, which is similar to expression (10) for a bridge-type switch.

If we use the parameters given above for the bridge-type-switch example, the value of $|\Delta r_k|$ for a differential switch as calculated from (20) will be equal to 24 ohm. However, such a large value of r_k is not characteristic of the

error introduced by the switch. In order to demonstrate this, let us examine in greater detail the range of application of (16) and (20) in a differential switch circuit.

The value of $r_k = 2r$ represents the constant resistance of the switch consisting of diodes with identical characteristics at a fixed temperature t and fixed values of controlling currents $I_{c2} = {}^1/_2 I_{c1}$. If for the same conditions of t = const and I_c = const we should take actual diode characteristics e'_n and r'_n which, owing to the dispersion of the parameters in a batch of diodes, will differ from the mean values e, r by the amounts of $\Delta e'_n$ and $\Delta r'_n$, the constant resistance of the switch will assume a new value

$$r'_k = r_k + (\Delta r_k)_{const}, \qquad (21)$$

where $(\Delta r_k)_{const}$ is determined from (20), since the signs of quantities $\Delta e'_1$ and $\Delta e'_2$ as well as of $\Delta r'_1$ and $\Delta r'_2$ can differ.

Let us now assume that conditions t = const and I_c = const no longer hold. Then another value of the key resistance $(\Delta r_k)_{var}$ is obtained, which must be calculated from (19), since the diode parameter additional variations $\Delta e''_1$ and $\Delta e''_2$ as well as $\Delta r''_1$ and $\Delta r''_2$ due to the effects of temperature and controlling currents have the same signs.

By comparing (19) and (20) it can easily be seen that $(\Delta r_k)_{var} \ll (\Delta r_k)_{const}$. Experiments have fully confirmed these conclusions. For a differential switch circuit with temperature variations in the range of + 10. . . +40°C and control current variations (owing to mains voltage changes) of ±10%, the variation of resistance $(\Delta r_k)_{var}$ is smaller by one order than the constant resistance component $(\Delta r_k)_{const}$, which in our example was equal to 24 ohm.

All the above reasoning fully applies to bridge-type switch circuits.

Thus, with respect to the error $\lambda = \dfrac{(\Delta r_k)_{var}}{R}$ introduced by a closed switch in the switched resistance (circuit shown in Fig. 4), both types of switches, the bridge and differential types, are equivalent to each other. But they differ in other respects.

In a bridge-type switch the actual constant resistance r'_k may amount to about 20 ohm, as has been shown in (9) and the cited example. In order to prevent this effect resulting in a noticeable error of the equivalent resistance shown in Fig. 4, it is necessary to reduce each of the switched resistors R by the value of r'_k, which is technically inconvenient.

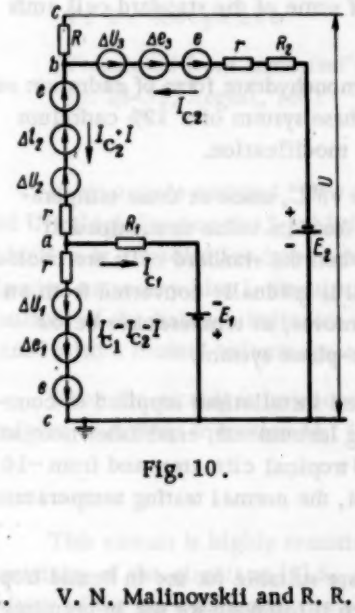

Fig. 10.

In a differential circuit, however, it is possible by adjusting the controlling currents to establish a condition for which the relationship $2r \approx -(\Delta r_k)_{const}$ holds. Then we shall find from (21) that $r'_k \approx 0$. Such an adjustment of the switching circuit can easily be checked by means of a millivoltmeter, which measures the voltage across the closed switch.

This constitutes the first advantage of a differential switch as compared with a bridge-type switch. It is true, that it is possible to adjust the bridge circuit in such a manner as to make points b-c (Fig. 5) equipotential under the effect of the controlling voltage E_c. For this purpose it is necessary to connect to the left-hand (or the right-hand) half of the bridge a rheostat with a slider at point a (or d). However, this operation leads to a considerable increase in the equivalent resistance of the switch. The second advantage of the differential switch consists in the fact that it is made up of 3 diodes and one controlling trigger, whereas a bridge-type switch is made up of 4 diodes and 2 controlling triggers.

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UTILIZATION OF UNSATURATED STANDARD CELLS UNDER SPECIAL CONDITIONS AND IN TROPICAL AND ARCTIC CLIMATES

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The operating GOST (All-Union State Standard) 1954-55 on standard cells (reference emfs) and instruction 176-55 for their checking completely satisfy the requirements of science and technology, providing the standard cells are used in the temperature interval of +10 to +40°C.

In connection with the extended range of application of standard cells there has arisen the necessity of widening their working temperature range.

In response to the requests made by our industry the standards of electrical units laboratory of the D. I. Mendeleev VNIIM (All-Union Scientific Research Institute of Metrology) has investigated the possibility of extending the range of unsaturated standard elements to 0 to +60°C, and of keeping them in a nonoperating conditions at temperatures as low as −40°C.

A large number of standard cells have been tested, both of old manufacture and those being produced at the present time by the "Teplokontrol'" plant (L'vov). On the basis of these experiments it is possible to arrive at the following conclusions.

Unsaturated standard cells containing a 12% cadmium amalgam can be used in automatic and electronic instruments in the temperature range of +5 to +55°C inclusive; moreover, the deviation of the actual value of the emf from that determined at 20°C will not exceed 100 μv, which meets the requirements of paragraph 4, GOST 1954-55.

The use of standard cells at higher temperatures is not recommended, since in such a case their emfs become unstable. For instance, in the temperature range of +55 to +60°C the deviations of some of the standard cell emfs from their value obtained at +20°C vary between +140 to −370 μv.

The reasons for such instability of the emfs consist of the conversion into a monohydrate form of cadmium sulfate crystal hydrates $3CdSO_4 \cdot 8H_2O$; the conversion at that temperature of the diphase system of a 12% cadmium amalgam into a liquid single-phase system and a possible change in the cadmium modification.

Neither is it recommended that standard cells be used at temperatures below +5°C, since at those temperatures their emf also becomes unstable. Moreover, in some of the cells it deviates from its value determined at 20°C by +0.1 to −0.1 mv. The large deviations of the emf are due the fact that when the standard cells are cooled below the temperature of +4°C cadmium sulfate crystals begin to form and the cell is gradually converted from an unsaturated into a saturated one rapidly changing its temperature coefficient. Moreover, at temperatures below +4°C the diphase system of 12% cadmium amalgam may change into a solid single-phase system.

The "general specifications for the manufacture of machines, instruments and installations supplied to countries with a tropical climate," approved by the Committee of Standards, Measuring Instruments, establishes nominal temperature ranges from 0 (minimum value) to +40°C (maximum value) for humid tropical climates, and from −10 to +55°C for dry tropical climates. According to the Indian Standard I. S. 589-1954, the normal testing temperature for instruments is fixed at 27 ± 5°C.

It will be seen from the above that standard cells made at the present time are suitable for use in humid tropical climates. For dry tropical climates they are only suitable for equipment which is intended for use in premises where the temperature does not change excessively.

Standard cells were tested in a nonoperating condition in a cryostat whose temperature was reduced gradually to −40°C. The electrolyte of a nonsaturated standard cell begins to freeze at a temperature of −15°C and it becomes completely solid at −23°C. The mean resistance of the unsaturated standard cells at a temperature of 20°C was of

the order of 400 ohm; when their temperature was lowered from 5 to 0°C the resistance rose to 750 ohm. For negative temperatures the resistance increases rapidly; thus, for −15.6°C it rises to 1250 ohm, for −23°C up to 4000 ohm, and for −40°C up to 60,000-70,000 ohm. According to paragraph 5 of GOST 1954-55 the internal resistance of newly manufactured unsaturated standard cells must not exceed 600 ohm.

The mean deviation of the actual emf of these standard cells from its value at 20°C decreases excessively (from −0.1 to −30 mv) in the temperature range of 0 to −40°C, and at −40°C the emf becomes equal to 1 v.

It will be seen from the above that at temperatures below +5°C standard cells sharply deviate from the requirements set by GOST 1954-55.

A complete restoration of the emf value to normal proceeds very slowly during one month. This is due to the fact that the solid phase of cadmium sulfate crystal hydrates formed during freezing transforms very slowly into a liquid phase which does not contain crystals at a temperature of +4°C.

The emf of standard cells returns to its normal value very slowly, after the cells have been subjected to a temperature of −40°C. In order to use standard cells at very low negative temperatures it is recommended that they be placed in thermostats which should prevent the freezing of the electrolyte, the dropping of their temperature below −15°C.

On the basis of the latest investigations carried out by the standards of electrical units laboratory of the VNIIM, it is advisable to introduce amendments in GOST 1954-55 which would provide the use of unsaturated standard cells in the temperature range of +5 to +55°C, thus improving the standard.

CERTAIN REMARKS ON D. KARO'S ARTICLE REGARDING THE NEW HIGH-PRECISION MEASURING CIRCUITS

V. P. Karpenko

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In his article entitled "New High-Precision Circuits for Measuring Impedances in the Range of Audio, Radio and Ultrahigh Frequencies," published in [1], D. Karo suggests several new circuits which provide measurements of inductive or capacitative impedances over a wide range of values at frequencies from 100 cps to 50 Mc. These circuits, named by him L-circuits, have been derived from a bridge circuit with differential transformers and the positions of the balance indicator and one bridge arm interchanged. Thus, an ordinary compensation circuit is obtained with a mutual inductance coupling between the primary and secondary circuits (Fig. 1).

When the circuit is balanced the voltage drop across impedance Z_2 is compensated by the secondary winding emf due to mutual induction M_2, i.e., the condition of balance is

$$U_{bc} + I_1 Z_2 = 0. \quad (1)$$

This circuit is highly sensitive. Assuming the balance indicator resistance to be infinite, we obtain the voltage sensitivity of the circuit as [2] $S_U = 1$ [v/v] and, hence, the total sensitivity of the circuit is determined by the sensitivity of the zero indicator, i.e., $S_k = S^{U_0}$ [mm/v].

The special value of such circuits, according to the author of the article, consists in their high precision (errors of the order of 0.001 and 0.01%), due to the fact that the common connection of the supply source and the balance indicator can be grounded, thus facilitating screening and the elimination of leakage currents at high frequencies. Moreover, D. Karo indicates the greater universality of the circuit in measuring impedances of any type and value.

Hence, the application of such circuits is very attractive owing to their simplicity, high sensitivity and their great accuracy as claimed by D. Karo.

Let us now examine whether in fact the above circuit possesses such universal characteristics and the possibility of measuring any impedances by replacing Z_1 and Z_2 with various resistances.

Let us represent the circuit in a general form (Fig. 2) and substitute mutual inductances M_1 and M_2 by voltages U_1 and U_2 which have random values, but are either in phase or 180° out of phase.

From Figs. 1 and 2 it follows that $\dot{U}_{ab} = \dot{U}_1$ and $\dot{U}_{bc} = \dot{U}_2$; hence, we obtain from (1)

$$\frac{\dot{U}_2}{\dot{U}_1} = -\frac{Z_3}{Z_1+Z_2}. \tag{2}$$

Let us find for what specific circuit parameters this balance condition holds.

Let us first examine a general case when both arms (Z_1 and Z_2) consist of impedances:

a) both of the same type (i.e., both inductive or capacitative)

$$Z_1 = r_1 \pm jx_1,$$
$$Z_2 = r_2 \pm jx_2.$$

By inserting these expressions in (2) we obtain the values of r_1 and x_1

$$r_1 = -\left(\frac{\dot{U}_1}{\dot{U}_2}+1\right) r_2, \tag{3a}$$

$$x_1 = -\left(\frac{\dot{U}_1}{\dot{U}_2}+1\right) x_2. \tag{3b}$$

These expressions show that in order to obtain a balance, vectors $\dot{U}_1$ and $\dot{U}_2$ should be in opposite directions and the inequality $U_1 > U_2$ should hold;

b) of different types (one capacitative and the other inductive)

$$Z_1 = r_1 \mp jx_1,$$
$$Z_2 = r_2 \pm jx_2.$$

By inserting these expressions into (2) we obtain the values of r_1 and x_1

$$r_1 = -\left(\frac{\dot{U}_1}{\dot{U}_2}+1\right) r_2, \tag{4a}$$

$$x_1 = \left(\frac{\dot{U}_1}{\dot{U}_2}+1\right) x_2. \tag{4b}$$

It will be seen from the above expressions that for such a condition the circuit will not balance, since in order to measure the resistivity component r_1 it is necessary to have vectors $\dot{U}_1$ and $\dot{U}_2$ in opposite directions, whereas for measuring the reactive component the two vectors $\dot{U}_1$ and $\dot{U}_2$ should be in the same direction.

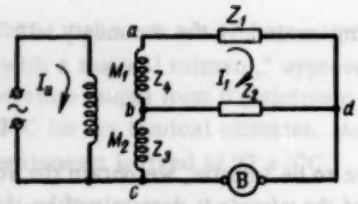

Fig. 1.

Let us examine a few particular cases when some of the impedance Z_1 and Z_2 components are equal to zero.

When Z_1 and Z_2 are purely resistive or reactive and of the same sign, i.e., when

$$Z_1 = r_1, \quad Z_2 = r_2, \quad \text{or} \quad Z_1 = \pm jx_1, \quad Z_2 = \pm jx_2,$$

they will balance according to equations (3a) and (3b), respectively.

When Z_1 and Z_2 are purely reactive, but with different signs

$$Z_1 = \mp jx_1,$$
$$Z_2 = \pm jx_2,$$

they will balance according to equation (4b).

When one of the arms is purely resistive and the other purely reactive, i.e., when

$$Z_1 = r_1, \quad Z_2 = \pm jx_2, \quad \text{or} \quad Z_1 = \pm jx_1, \quad Z_2 = r_2,$$

it will be impossible to balance the circuit.

When actual circuits are examined, i.e., when the effect of impedance Z_4 of the secondary winding which has a mutual inductance M_1 (Fig. 1) is taken into account, we obtain a slightly more complicated balance condition:

$$\frac{M_2}{M_1} = -\frac{Z_2}{Z_1 + Z_2 + Z_3} \tag{5}$$

By analyzing (5) and examining circuits with different types of impedances Z_1 and Z_2, in a manner similar to the above, we come to the following conclusions.

When impedances Z_1 and Z_2 are complex and of a similar type their balance equations will be:

$$r_1 = -\left(\frac{M_1}{M_2} + 1\right) r_2 - r_4, \tag{6a}$$

$$x_1 = -\left(\frac{M_1}{M_2} + 1\right) x_2 - x_4. \tag{6b}$$

It will be seen from these equations that the circuit will balance if the mutual inductance windings M_1 and M_2 are connected in opposition. Moreover, the following inequality must hold:

$$\left(\frac{M_1}{M_2} - 1\right) r_2 > r_4, \tag{7}$$

and for inductive impedances Z_1 and Z_2 the following equality must also hold:

$$\left(\frac{M_1}{M_2} - 1\right) x_2 > x_4. \tag{8}$$

When impedances Z_1 and Z_2 are capacitative, the minimum measurable reactance x_1 will be limited by the constant inductive reactance x_4.

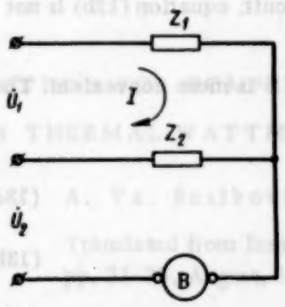

Fig. 2.

When the reactances of impedances Z_1 and Z_2 have opposite signs their balance equations will be

$$r_1 = -\left(\frac{M_1}{M_2} + 1\right) r_2 - r_4, \tag{9a}$$

$$x_1 = \left(\frac{M_1}{M_2} + 1\right) x_2 \mp x_4. \tag{9b}$$

When Z_1 is inductive and Z_2 is capacitative, the circuit will not balance, since in order to measure the resistive component r_1 it is necessary to have the mutual inductance M_1 and M_2 windings connected in opposition, whereas in order to measure the reactive component x_1 it is necessary to have them in series-aiding.

When Z_1 is capacitative and Z_2 is inductive the circuit will only balance if

$$x_4 > \left(\frac{M_1}{M_2} - 1\right) x_2,$$

which is hardly convenient for measurements, since the maximum measured reactance is limited by the value of x_4.

and x_2 must vary between 0 and x_4 for variations of x_1 between x_4 and 0, respectively. Thus, the circuit is only suitable for measuring large values of capacitance C_1, which is seldom required at high frequencies.

When impedances Z_1 and Z_2 are purely resistive or purely reactive, the circuit will not balance since it will then be necessary to have $x_4 = 0$ or $r_4 = 0$, respectively, which cannot be obtained in practice.

When one of the impedances is purely resistive and the other purely reactive, circuits are obtained which cannot be used for measuring purposes.

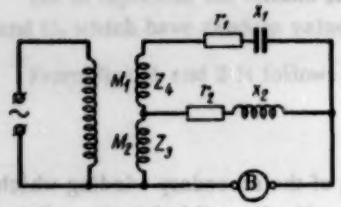

Fig. 3.

Thus, when $Z_1 = r_1$ and $Z_2 = \pm jx_2$ the balancing conditions will be

$$r_1 = -r_4, \tag{11a}$$

$$x_2 = \frac{x_4}{\frac{M_1}{M_2} + 1}. \tag{11b}$$

Such a circuit cannot be balanced at all.

For $Z_1 = \pm jx_1$ and $Z_2 = r_2$ the balance conditions are

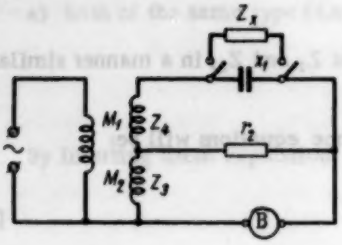

Fig. 4.

$$x_1 = x_4, \tag{12a}$$

$$r_2 = -\frac{r_4}{\frac{M_1}{M_2} + 1}. \tag{12b}$$

Such a circuit can only be used for measuring resistance r_2 with a variable ratio between the mutual inductances when they are connected in opposition. The capacitative reactance x_1 cannot be measured at all, since x_4 is constant.

On the basis of the above considerations let us examine some of the circuits suggested by D. Karo.

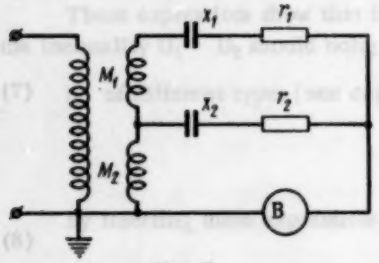

Fig. 5.

The circuit shown in Fig. 3 is intended for measuring a capacitative impedance. Its balance conditions are given in (9a) and (9b) and it can only be used for measurements if it meets condition (10). It will be seen that this circuit is far from being universally applicable.

The circuit shown in Fig. 4 is also intended for measuring capacitance. The circuit is balanced twice, the first time without the tested impedance, and the second time with impedance Z_X connection.

The balance conditions for this circuit with Z_X disconnected are obtained from (12a) and (12b). It will be seen from the above equations that the circuit can only be balanced for $1/\omega C_1 = x_4$. By connecting $Z_X = r_X - jx_{C_X}$ to the circuit, equation (12b) is not changed and the circuit cannot be used for measuring capacitance, since x_4 is constant.

It seems to us that for measuring capacitative impedance the circuit shown in Fig. 5 is more convenient. The balance conditions for this circuit are:

$$r_1 = -\left(\frac{M_1}{M_2} + 1\right) r_2 - r_4, \tag{13a}$$

$$x_1 = -\left(\frac{M_1}{M_2} + 1\right) x_2 + x_4. \tag{13b}$$

Let us now examine briefly the accuracy of the above circuits.

In the first place the circuit balance conditions include impedance Z_4 of the secondary winding with mutual inductance M_1. This impedance affects the measurement results and an accurate evaluation of this effect complicates the operation of the circuit. Moreover, the stray capacitances C'_4 (Fig. 6) which shunt the winding will distort impedance Z_4 in a random manner. An accurate estimation or elimination of these capacitances is impossible. Capacitance C'_1, which shunts impedance Z_1, and capacitance C'_2, which shunts impedance Z_2, will also affect the accuracy of measurements. The effect of these capacitances is also difficult to account for or eliminate.

Capacitances C''_2 and C'_3 may produce phase differences between the emf $j\omega M_2 I_U$ (Fig. 1) and the voltage drop $I_1 Z_2$; this can occur especially at high frequencies when the capacitive reactances become of the same order as ωM_2.

Even a most approximate analysis shows that the effect of the above stray capacitances on high-frequency measurements will produce a considerable error.

This circuit could be used more successfully, in our opinion, for instance, for mass-production testing of mutual inductance coils by connecting them instead of M_2 and using reference coils in place of M_1 (with identical primary windings for the measured and reference coils). This measuring method can be made sufficiently accurate, since it is possible to eliminate the effect of all the stray capacitances by using the substitution technique. In this case with unknown impedances Z_1, Z_2, and Z_4 the mutual inductance M_2 can easily be calculated from

$$M_2 = -M_1 K, \qquad (14)$$

where $K = \dfrac{Z_2}{Z_1 + Z_2 + Z_4}$ is determined from measurements with reference M_1 and M_2.

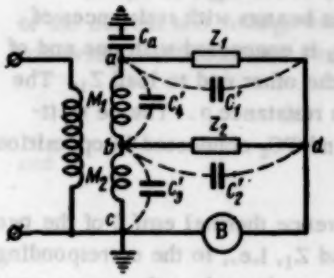

Fig. 6.

This method is advantageous because its measurement results are not affected by the impedance of the tested coil.

SUMMARY

1. The circuits suggested by D. Karo deserve consideration for their simplicity and high sensitivity and can be used successfully for many measurements.

2. The accuracy of measurements of these circuits does not exceed the ordinary compensation circuits.

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METHOD FOR COMPENSATING POWER CONSUMPTION IN THERMAL WATTMETERS

A. Ya. Besikovich and D. I. Zorin

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It is known that in order to avoid errors due to the power consumption of both electrodynamic and thermoelectric wattmeters when they are used for measuring power in a load, it is necessary to apply corrections whose calculation entails the knowledge of the wattmeter circuit current, voltage and impedances.

The development of a wattmeter which avoid this drawback by means of a special circuit is undoubtedly of interest. Several thermal-wattmeter circuits which have solved this problem have been analyzed by Bader [1]. However, Bader's condition that the two thermal converters used in the circuits must have completely identical square-

law volt-ampere characteristics could not be attained in practice. This explains why up to the present time there are no instruments which do not require corrections of their readings.

In the circuit proposed by us it is sufficient to have converters with square-law volt-ampere characteristics and it is not necessary that the two characteristics should be identical. Moreover, Bader's circuits did not provide a higher accuracy for thermoelectric wattmeters; neither did they reduce appreciably their temperature error. Both these problems are solved in the instrument designed by us.

Figure 1 shows a schematic of a thermoelectric wattmeter for measuring power in load Z_1 which takes a current I and provides a voltage drop U (their instantaneous values are equal to $\underline{i}$ and $\underline{u}$, respectively).

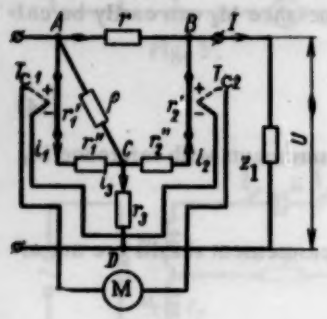

Fig. 1.

Instantaneous current $\underline{i}$ flows through wattmeter shunt $\underline{r}$ and a circuit in parallel with it consisting of two thermal-converter heaters with resistances of r'_1 and r'_2, respectively. A multiplier resistance r_3 is connected with one end of the junction point C of the two heaters, and with the other end to load Z_1. The heater circuit of the first converter is shunted with resistance ρ. The dc wattmeter circuit consists of two thermocouples TC_1 and TC_2 connected in opposition and a moving coil meter M.

Let us find the condition for which the difference thermal emf E of the two converters will be proportional to the power in load Z_1, i.e., to the corresponding values of U, I and the circuit parameters, and will not depend on the power consumption of the wattmeter. Providing the volt-ampere characteristics of the thermal converters follow a square law, the thermal emf can be represented in terms of heater currents i_1 and i_2 by the expression

$$E=E_1-E_2=\frac{k_1}{T}\int_0^T i_1^2dt-\frac{k_2}{T}\int_0^T i_2^2dt, \tag{1}$$

where k_1 and k_2 are the coefficients of proportionality which characterize the sensitivity of thermocouples.

Let us express the values of currents i_1 and i_2 by means of current $\underline{i}$, voltage $\underline{u}$ and the circuit parameters. For this purpose let us assume that $r_1 = R'_1 + r''_1$ and $r_2 = r'_2 + r''_2$ and solve two simultaneous equations obtained from Fig. 1:

$$u=i_3r_3-i_2r_2 \tag{2}$$

$$ir=i_1r_1+i_2(r_2+r). \tag{3}$$

Since:

$$i_3=i_1\left(1+\frac{r_1}{\varrho}\right)-i_2, \tag{4}$$

we find from (2) and (3) that:

$$i_1=\frac{u(r+r_3)+ir+(r_2+r_3)}{r_3\left(1+\frac{r_1}{\varrho}\right)(r+r_2)+r_1(r_2+r_3)}; \tag{5}$$

$$i_2=\frac{irr_3\left(1+\frac{r_1}{\varrho}\right)-ur_1}{r_3\left(1+\frac{r_1}{\varrho}\right)(r+r_2)+r_1(r_2+r_3)}. \tag{6}$$

By inserting the current values found from (1) we obtain

$$E=\frac{1}{\left[r_3\left(1+\frac{r_1}{\varrho}\right)(r+r_2)+r_1(r_2+r_3)\right]^2}\cdot\frac{1}{T}\int_0^T\Big[k_1u^2(r+r_2)^2+k_12ui(r+r_2)r(r_2+r_3)+k_1i^2r^2(r_2+r_3)^2-k_2i^2r^2r_3\left(1+\frac{r_1}{\varrho}\right)^2+$$
$$+k_22uirr_3\left(1+\frac{r_1}{\varrho}\right)-k_2u^2r_1^2\Big]dt.$$

Assuming for simplicity of mathematical calculations that voltage u and current i are sinusoidal and their phase difference is equal to φ, we have

$$E=\frac{1}{\left[r_3\left(1+\frac{r_1}{\varrho}\right)(r+r_2)+r_1(r_3+r_2)\right]^2}\left[k_1U^2(r+r_2)^2+k_1 2UI\cos\varphi(r+r_2)r(r_3+r_2)+k_1I^2r^2(r_3+r_2)^2-\right.$$
$$\left.-k_2I^2r^2r_3^2\left(1+\frac{r_1}{\varrho}\right)^2+k_2 2UI\cos\varphi r r_1 r_3\left(1+\frac{r_1}{\varrho}\right)-k_2U^2r_1^2\right]. \tag{7}$$

It will be seen from (7) that the difference thermal emf is proportional not only to the power, but to the squares of the current and voltage values. In order to eliminate the effect of the last two parameters on the value of E the following conditions must be fulfilled:

$$r_1=\sqrt{\frac{k_1}{k_2}}(r+r_2) \tag{8}$$

and

$$\left(1+\frac{r_1}{\varrho}\right)=\sqrt{\frac{k_1}{k_2}}\left(1+\frac{r_2}{r_3}\right). \tag{9}$$

Then expression (7) assumes the form

$$E=k_2\frac{r}{r+r_2}\cdot\frac{1}{r_3+r_2}UI\cos\varphi, \tag{10}$$

which shows that the difference thermal-emf is proportional only to the measured power. The coefficient of proportionality $k_2\frac{r}{r+r_2}\cdot\frac{1}{r_3+r_2}$ depends on the sensitivity of the thermal converter k_2, the value of shunt resistance r, the multiplying resistance r_3, and the second converter heater circuit resistance r_2.

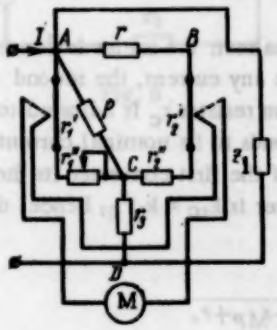

Fig. 2.

Since (10) only holds when conditions (8) and (9) are fulfilled, it is necessary to be able to provide and control these conditions. It is possible to obtain this automatically by means of two operations which are known as voltage and current balancing, respectively, if resistance ρ and resistance r_1 in the first converter heater circuit are made variable.

Voltage balancing, i.e., obtaining the equality of $E_1 = E_2$ is attained by connecting the thermal wattmeter as shown in Fig. 2. This connection differs from Fig. 1 by the fact that the load current does not flow through the shunt, since the load is disconnected from point B. In this circuit shunt r and the second converter heater circuit which is in series with the shunt are connected in parallel with the first converter heater circuit. Thus, the currents I_1 and I_2 in the converter heaters will only depend on voltage U_1 across AC which is proportional to the total voltage

$$I_1=\frac{U_1}{r_1} \quad \text{and} \quad I_2=\frac{U_1}{r+r_2}.$$

Hence

$$E_1=k_1I_1^2=k_1\frac{U_1^2}{r_1^2}; \qquad E_2=k_2I_2^2=k_2\frac{U_1^2}{(r+r_2)^2}.$$

Voltage balancing, hence the equality of $E_1 = E_2$ (no reading on the meter), is attained when

$$r_1=\sqrt{\frac{k_1}{k_2}}(r+r_2),$$

i.e., when condition (8) is fulfilled. Since k_2, r and r_2 also figure in equation (10) they must not be changed in order

to keep the coefficient of proportionality constant. Hence, for any value of k_1 the equality $E_1 = E_2$ can be attained by varying the resistance in the first heater circuit. For this purpose variable resistor r''_1, shown in Fig. 2, is connected in series with the heater.

The thermal-wattmeter current balancing is carried out by means of the circuit shown in Fig. 3. The multiplying resistor r_3 is disconnected from point D and connected to point B. The current then passing through the heaters will not depend on the voltage applied to load Z_1, but will only depend on the current flowing through the load. If an equality of $E_1 = E_2$ is attained in a manner similar to voltage balancing we shall obviously have:

$$k_1 I_1^2 = k_2 I_2^2 \text{ and } \frac{I_2}{I_1} = \sqrt{\frac{k_1}{k_2}}. \tag{11}$$

From the circuit in Fig. 3 we obtain:

$$I_1\left(1+\frac{r_1}{\varrho}\right) = I_2\left(1+\frac{r_2}{r_3}\right),$$

or taking (11) into account we have

$$1+\frac{r_1}{\varrho} = \sqrt{\frac{k_1}{k_2}}\left(1+\frac{r_2}{r_3}\right),$$

which coincides with (9). For the purpose of this adjustment of the circuit, resistor ρ is made variable.

However, the values of r_2 and k_2 figuring in (10) can also change with time and due to varying conditions of operation. In order to prevent the variations in r_2 affecting the instrument readings an adjustable resistor r''_2 is provided as shown in Fig. 3. The effect of changes in k_2 is eliminated by means of a special calibration method which consists in setting the calibration current I_C and then calibrating the thermal wattmeter.

A dc calibration current I_C is set by means of the circuit shown in Fig. 4. The heater of the adjusted (second) thermal converter is connected in series with a dc source. Current I_C is set by means of variable resistor r_p, and measured by means of the thermal-wattmeter 0.1 ohm meter used with shunt r_s and building-out resistor r_b. The value of r_b is selected to provide a full-scale meter deflection.

The calibration of the thermal wattmeter proper is carried out in the following manner: the meter is connected in series with the two thermocouples (Fig. 5) (the first converter heater does not take any current, the second heater takes the calibration current I_C), the calibration resistor r_c is adjusted to obtain a maximum reading on the meter which corresponds to its nominal current I_n. Since there is no current flowing through the heater of the first converter its thermal emf is $E_1 = 0$, and the thermal emf of the second heater is $E_{2C} = k_2 I_C^2$; hence, the current in the thermocouple circuit is

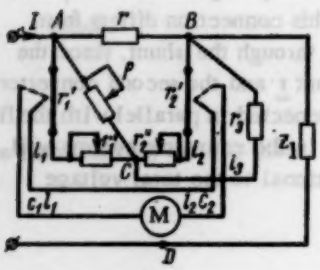

Fig. 3.

$$I_K = \frac{E_{2K}}{r_{TC1}+r_{TC2}+r_{MP}+r_K} = \frac{k_2 I_K^2}{r_{TC1}+r_{TC2}+r_{MP}+r_K}.$$

where r_M is the resistance of the meter; r_{TC1} and r_{TC2} are the resistances of the first and second thermocouples.

Thus, by adjusting resistor r_c the expression

$$\frac{k_2}{r_{TC1}+r_{TC2}+r_{MP}+r_K} = \frac{I_C}{I_K^2} \tag{12}$$

is maintained at a constant value irrespective of variations in k_2, r_{TC1} and r_{TC2}.

A thermal wattmeter which has a circuit as described above is not calibrated in terms of power, but in terms of the calibrating current, since this current in the thermocouple circuit is equal to

$$I_T = \frac{E}{r_{TC1}+r_{TC2}+r_{MP}+r_K} = \frac{k_2}{r_{TC1}+r_{TC2}+r_{MP}+r_K} \cdot \frac{r}{r+r_3} \cdot \frac{UI\cos\varphi}{r_3+r_3}.$$

Having carried out the above calibrations we shall obtain:

$$I_T = \frac{I_n}{I_c^2} \cdot \frac{r}{r+r_2} \cdot \frac{1}{r_3+r_3} UI\cos\varphi. \qquad (13)$$

Hence, the current in the thermocouple circuit is proportional to the measured power. We thus obtain a single-value determination of the coefficient of proportionality by means of resistors r, r_2 and r_3 and currents I_c and I_n. Thus, the thermal wattmeter need not be calibrated in terms of power, and the correctness and accuracy of its measurements is determined by the accuracy of setting its resistors and measuring currents I_c and I_n. It is obvious that if the instrument is designed to measure at several power factors, the calibrating current must be adjusted for each of them.

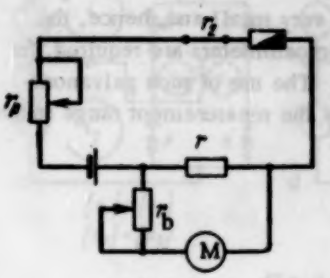

Fig. 4.

A complete circuit of a multirange thermal wattmeter which is provided with facilities for all the required balancing and calibrations is shown in Fig. 6. Switches S_1 and S_2 serve to set the current and voltage ranges respectively. Switch S_3 serves to set the circuit for the required balancing and measuring operations. In position 1 (measurement) the circuit shown in Fig. 1 is provided; in position 2 the circuit shown in Fig. 2, and in position 3 (current balancing) the circuit shown in Fig. 3.

The balancing is carried out by means of resistors r_1-r_2 (for current) and r_3-r_4 (for voltage).

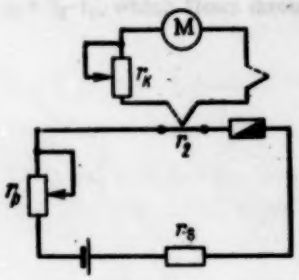

Fig. 5.

Switch S_4 serves to set the circuits for measurement (position 1), calibration (position 2), and the setting of the calibrating current (position 3). In the last two instances the thermocouple heaters are disconnected from their shunts and building out resistors and connected to the auxiliary dc circuit. In position 3 the circuit shown in Fig. 4 is obtained and in position 2 the circuit shown in Fig. 5.

Switch S_5 serves to vary the nominal power factor of the instrument by means of changing the calibration current.

This thermal wattmeter can be used not only to measure power across a load, but also at the output terminals of a generator. For this purpose, it is only necessary to transfer the generator terminal from point A to point B (Fig. 1) and reverse the polarity of the meter. By providing this facility a universal wattmeter is obtained. By using the same method as that used for analyzing the circuit in Fig. 1 we obtain the same equation (10) providing the conditions given in (8) and (9) are maintained.

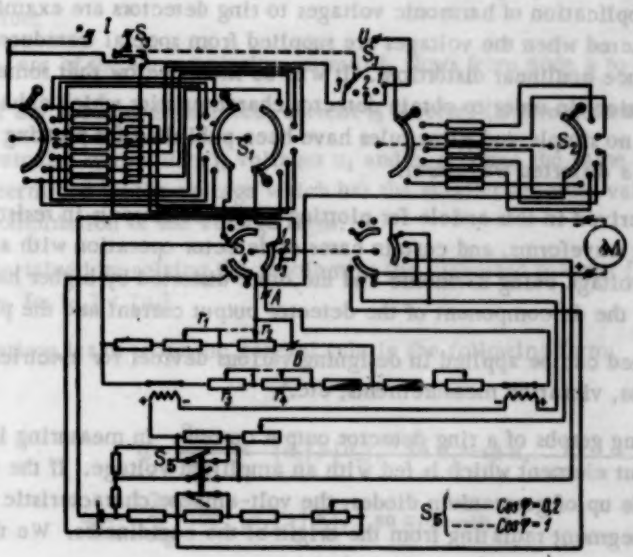

Fig. 6.

Several thermal wattmeters have been produced in the VNIIM (All-Union Scientific Research Institute of Metrology) according to the circuit given in Fig. 6 and provided with 7 current measuring ranges from 0.1 to 10 amp, 6 voltage measuring ranges from 5 to 200 v, and two nominal power factors of 0.2 and 1. For measuring purposes the instrument uses a mass-produced Soviet-made dc microammeter type M95 on its 10 μa range. The accuracy of the wattmeter readings is ascertained by checking the calibration current, the shunt and building-out resistors and the resistance of the second converter heater.

The wattmeter was tested on a direct current by means of 2 potentiometers with separate supply circuits. The tests have shown that the referred error of the thermal wattmeter is equal to the error of the microammeter and does not exceed 1%.

The wattmeter uses multielement thermal converters [2] which, owing to their design and method of operation, provide a square-law volt-ampere characteristic. If thermal converters type TVB, which are mass-produced by our industry, are used the thermal-emf at the square-law portion of their characteristic is very small and, hence, the thermocouple circuit current is also small. In order to measure it highly sensitive microammeters are required, for instance, photoamplifier galvanometers type F-116/2 made by the "Vibrator" plant. The use of such galvanometers will obviously reduce the measurement accuracy, but it will extend considerably the measurement range in the direction of lower power, which, in many instances is very important.

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2. A. Ya. Bezikovich and D. I. Zorin, Trudy VNIIM 1958, 28/88.

OPERATION OF A RING DETECTOR WITH AN APPLIED VOLTAGE OF A DISTORTED WAVEFORM

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Certain cases of the application of harmonic voltages to ring detectors are examined in [1-4]. In practice other waveforms are encountered when the voltages are supplied from special transducers, stabilizers, limiting or other elements which introduce nonlinear distortions. It will be shown below that sometimes it is advantageous to deliberately introduce distortions in order to obtain detector characteristics which enhance the stability of the measuring instrument. However, no simple and clear rules have been published for plotting graphs of detector output currents for applied voltages of a distorted waveform.

Formulas have been derived in this article for plotting graphs of currents in resistive loads of ring detectors with arbitrary supply voltage waveforms, and certain cases of detector operation with a purely harmonic and periodic voltage are examined, one voltage being harmonic and the other distorted by higher harmonics. The effect of distortion on the relation between the dc component of the detector output current and the phase angle are also determined.

The results thus obtained can be applied in designing various devices for electrical measurements of mechanical parameters (strain gauges, vibration measurements, etc.).

General rule for plotting graphs of a ring detector output current. In measuring instruments the ring detector is used, as a rule, as an output element which is fed with an amplified voltage. If the voltage level is sufficiently high and the detector is made up of germanium diodes, the volt-ampere characteristic of each rectifier can be represented by a straight line segment radiating from the origin of the coordinates. We therefore assume the forward diode resistance to be R and its reverse resistance equal to ∞.

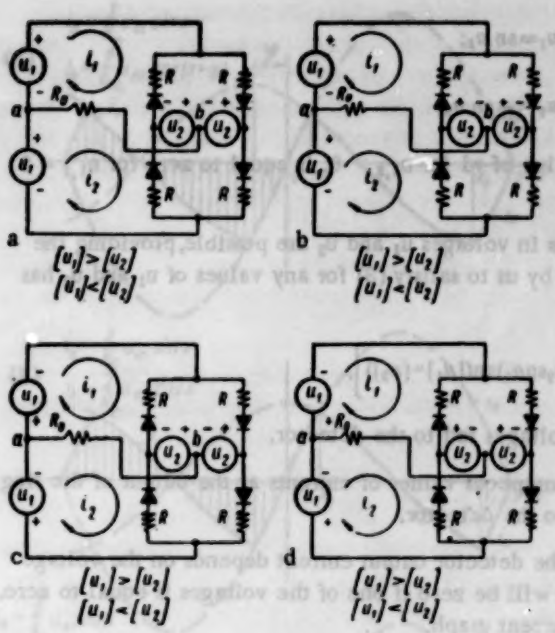

Fig. 1.

The load impedance R_0 is considered to be purely resistive, which is the case if the ring detector is connected to a moving coil meter or any other inertial instrument, for instance, an integrating loop oscilloscope.

Let us analyze a ring detector circuit (Fig. 1) for various combinations of supply voltages with different magnitudes and signs. Let us stress that we shall first examine instantaneous (not amplitude) values of voltages and currents in the detector circuit. It is obvious that for two voltages u_1 and u_2 the number of possible sign combinations is equal to 4 for $[u_1] > [u_2]$, it is also equal to 4 for $[u_1] < [u_2]$.

The total number of possible combinations amounts to 8.

By means of Kirchhoff's law we can obtain equations for currents i_1 and i_2 which flow through the detector circuit when $[u_1] > [u_2]$ (Fig. 1) for each of the possible combinations. For instance, for a combination shown in Fig. 1a the equations have the form:

$$\begin{aligned} i_1(R+R_0)-i_2R_0&=u_1-u_2;\\ i_2(R+R_0)-i_1R_0&=u_1+u_2. \end{aligned} \qquad (1)$$

By solving (1) and the equations obtained for the remaining combinations (Figs. 1b, 1d, 1c), with respect to current $i_0 = i_2 - i_1$, which flows through load R_0 we obtain, respectively:

$$\begin{aligned} &1)\ i_0=\frac{2u_2}{R+2R_0};\\ &2)\ i_0=-\frac{2u_2}{R+2R_0};\\ &3)\ i_0=-\frac{2u_2}{R+2R_0};\\ &4)\ i_0=\frac{2u_2}{R+2R_0}. \end{aligned} \qquad (2)$$

From these equations we can assert that:

1) if $[u_1] > [u_2]$, the absolute value of the output detector current is proportional to $[u_2]$, i.e., to the voltage with the smaller absolute value;

2) if voltages u_1 and u_2 are of the same polarity, current i_0 flows from node a to node b (Fig. 1);

3) if the sign of one of the voltages is reversed, current i_0 reverses its direction and flows from b to a.

Providing the circuit remains symmetrical, voltages u_1 and u_2 produce the same effects and, therefore, when $[u_1] < [u_2]$, current $[i_0]$ is determined by the voltage which has the smaller absolute value, and the sign of the current will be determined by the combination of the voltage signs.

This can easily be ascertained by solving the problem in the suggested manner for each of the possible combinations of signs in u_1 and u_2 for $[u_1] < [u_2]$.

Without further evaluations let us write the general rule in the following form:

$$\begin{aligned} i_0&=\frac{2u_2}{R_e} \text{ for } [u_1]>[u_2],\ \operatorname{sn} u_1=\operatorname{sn} u_2;\\ i_0&=-\frac{2u_2}{R_e} \text{ for } [u_1]>[u_2],\ \operatorname{sn} u_1=-\operatorname{sn} u_2; \end{aligned} \qquad (3)$$

$$i_0=\frac{u_1}{R_e} \text{ for } [u_1]<[u_2],\ \operatorname{sn} u_1=\operatorname{sn} u_2;$$
$$i_0=-\frac{2u_1}{R_e} \text{ for } [u_1]<[u_2],\ \operatorname{sn} u_1=-\operatorname{sn} u_2, \tag{3}$$

where sn $u_{1,2}$ is the sign of $u_{1,2}$, i.e., a function which has the value of +1 for $u_{1,2} > 0$, is equal to zero for $u_{1,2} = 0$ and equal to −1 for $u_{1,2} < 0$, and $R_e = R + 2R_0$.

All the above combinations of the absolute values and signs in voltages u_1 and u_2 are possible, providing the voltages change according to a random law. The function found by us to satisfy (3) for any values of u_1 and u_2 has the form:

$$i_0=\frac{1}{R_e}\left[u_1\operatorname{sn}u_2+u_2\operatorname{sn}u_1-(u_1\operatorname{sn}u_2-u_2\operatorname{sn}u_1)\operatorname{sn}([u_1]-[u_2])\right], \tag{4}$$

where $[u_1]$ and $[u_2]$ are the absolute values of the instantaneous voltages fed to the detector.

Expression (4) is a design formula for determining the instantaneous values of currents at the output of the ring detector for random variations of the instantaneous voltages fed to the detector.

It will be seen from (3) and (4) that the absolute value of the detector output current depends on the voltage which has the smaller absolute value. Hence, the output current will be zero if one of the voltages is equal to zero. We use this rule subsequently in plotting a ring detector output current graph.

Response of the detector to purely harmonic voltages. Let us examine the first particular case of supplying the detector with voltages which vary according to the law

$$u_1=u_{01}\sin x,$$
$$u_2=u_{02}\cos(x+\varphi), \tag{5}$$

where u_1 and u_2 are the instantaneous values of the voltages; u_{01} and u_{02} are the amplitudes of these voltages.

Such a phase relation between voltages u_1 and u_2 exists when the detector is used as an FM signal demodulator.

Figure 2 shows the graphs plotted according to (4). In these graphs i_0 is the detector output current. Let us determine the dc component of current i_0 as a function of amplitudes u_{01} and u_{02} and of the phase shift angle φ of one of the voltages.

For this purpose it is sufficient to evaluate the integral of current i_0 for half the period of one of the input voltages

$$I_0=\frac{1}{\pi}\int_0^{\pi} i_0\,dx. \tag{6}$$

According to (4) and (5) and Fig. 2, integral (6) is equal to the difference of the shaded areas, in other words to the sum of the integrals of separate sections

$$I_0=\frac{2}{\pi R_e}\left[u_{01}\int_0^{x_1}\sin x\,dx+u_{02}\int_{x_1}^{x_2}\cos(x+\varphi)\,dx-u_{02}\int_{x_2}^{\pi}\sin x\,dx\right]; \tag{7}$$

$$I_0=\frac{2}{\pi R_e}\left\{u_{02}[\sin(x_2+\varphi)-\sin(x_1+\varphi)]-u_{01}(\cos x_1+\cos x_2)\right\}=\frac{4u_{02}}{\pi R_e}F(\varphi).$$

Angles x_1 and x_2 are determined from condition

$$u_1(x_1)=u_2(x_1);$$
$$u_1(x_2)+u_2(x_2)=0. \tag{8}$$

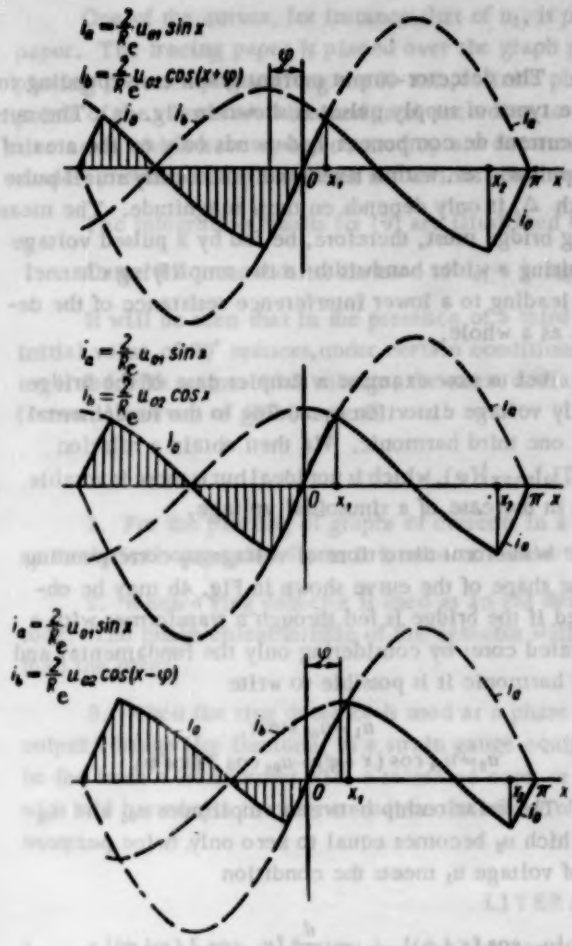

Fig. 2.

For the values of u_1 and u_2 from (5) we have

$$x_1 = \text{arc tg} \frac{\cos\varphi}{\frac{u_{01}}{u_{02}} + \sin\varphi},$$

$$x_2 = \text{arc tg} \frac{\cos\varphi}{\sin\varphi - \frac{u_{01}}{u_{02}}}.$$

Figure 3 shows the design graphs of the relationship

$$F(\varphi) = \frac{I_0}{[I_0]_{\max}} =$$

$$= \frac{1}{2}\left[\sin(x_2+\varphi) - \sin(x_1+\varphi) - \frac{u_{01}}{u_{02}}(\cos x_1 + \cos x_2)\right]$$

for the two cases:

$$\frac{u_{01}}{u_{02}} \gg 1, \quad \frac{u_{01}}{u_{02}} = 1 \text{ (curves 1 and 2).}$$

In the first case $F(\varphi) \approx -\sin\varphi$ and the detector becomes a linear vectormeter, i.e., a device whose output current is proportional to the projection of one vector $\hat{u}_2$ onto the other $\hat{u}_1$ [2, 4].

For $u_{01} = u_{02}$ the characteristic of $F(\varphi)$ has a very large linear section (curve 2 in Fig. 3) and the detector can be used as a phasemeter in the range of +90 to −90°.

Effect of higher harmonics on the relation between the output current dc component and angle φ. Let us now examine the operation of a ring detector as the output element of an instrument which measures the voltage across the diagonal of an ac bridge.

It is known that, providing the bridge consists of resistances, the ring detector will receive two voltages (one from the measuring diagonal of the bridge through an amplifying channel, and the other from the generator which feeds the bridge) which will be either in or out of phase, $\varphi = +90°$ or $\varphi = -90°$ (see Eq. 5).

Thus, the detector output current assumes, as will be seen from curves 1 and 2 in Fig. 3, either a negative or a positive value corresponding to angle $\pm 90°$ and depending on the sign of the unbalance. The absolute value of the current should then depend on the bridge output voltage only.

It is also known that, owing to stray capacitances in the bridge (with long supply leads), the bridge output voltage is slightly displaced in phase with respect to the supply voltage and, hence, strictly speaking, $\varphi \neq \pm 90°$.

If the stray capacitances are unstable (for instance, in strain gauge measurements of a moving object), the detector output current will depend not only on the amplitude of the output current in the bridge, but also on the instability of the phase difference between voltages u_1 and u_2. If the characteristic $F(\varphi)$ has the shape of curves 1 and 2 (Fig. 3), large measurement errors will be incurred.

It is obvious that for measuring the active component of the bridge output voltage it is advantageous to have the characteristic representing the relation between the dc component of the detector output current and the phase an angle of the shape of an open polygon ABCODEM (Fig. 3).

With such a relationship even considerable deviations of angle φ from $\pm 90°$ will not affect the measurement results. The idealized characteristic of the shape of ABCODEM corresponds to the case when the detector is fed with a harmonic voltage u_1, whose amplitude $u_{01} \to \infty$, and the other voltage u_2 has the form of π-shaped pulses with alternating signs, whose repetition frequency equals the frequency of u_1, and the duration of each pulse $\Delta \to 0$.

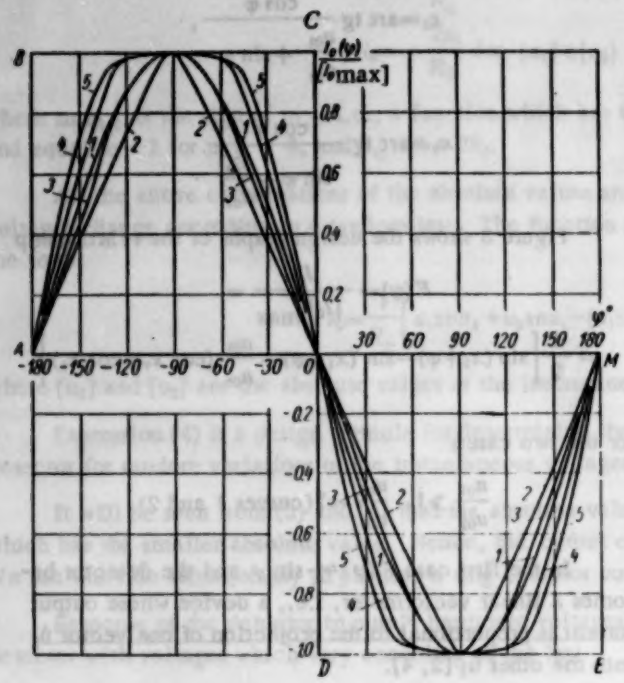

Fig. 3. Ring detector characteristics for certain types of voltages u_1 and u_2 applied to it. $I_0(\varphi)$ is the dc component of the tector output current; $[I_0 \max]$ is the absolute maximum value of the dc component for $\varphi = \pm 90°$.

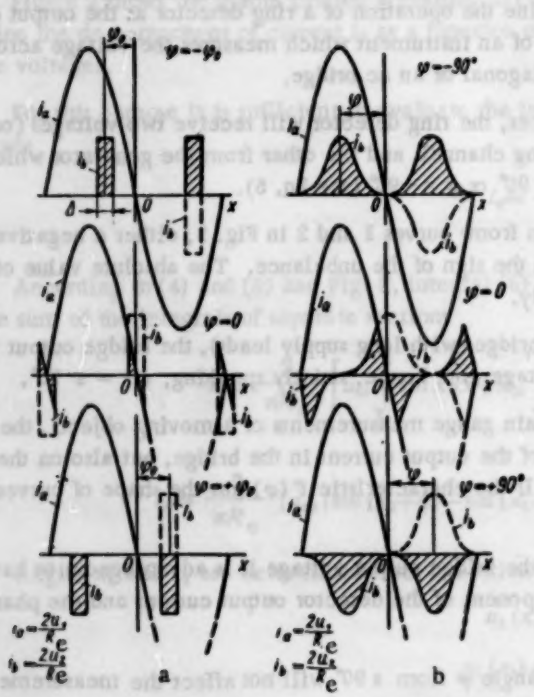

Fig. 4.

The detector output current graph corresponding to three types of supply pulses is shown in Fig. 4a. The output current dc component I_0 depends only on the area of the pulses, i.e., with a fixed and sufficiently small pulse width Δ, it only depends on their magnitude. The measuring bridge must, therefore, be fed by a pulsed voltage requiring a wider bandwidth in the amplifying channel and leading to a lower interference resistance of the device as a whole.

Let us now examine a simpler case of the bridge supply voltage distortion by adding to the fundamental only one third harmonic. We then obtain a relation $\{I_0/[I_0]_{max}\}(\varphi)$, which is not ideal but is more favorable than in the case of a sinusoidal voltage.

Waveform distortions of voltage u_2 corresponding to the shape of the curve shown in Fig. 4b may be obtained if the bridge is fed through a transformer with a saturated core; by considering only the fundamental and third harmonic it is possible to write

$$u_1 = u_{01} \sin x,$$
$$u_2 = u_{04} \cos(x+\varphi) - u_{05} \cos 3(x+\varphi).$$

The relationship between amplitudes u_{04} and u_{05} for which u_2 becomes equal to zero only twice per period of voltage u_1 meets the condition

$$-\frac{d}{dx}[u_{04}\cos(x+\varphi)]_{x=x_0} = \frac{d}{dx}[u_{05}\cos 3(x+\varphi)]_{x=x_0},$$

$$x_0 + \varphi = \frac{\pi}{2},$$

from which we obtain

$$\frac{u_{05}}{u_{04}} = \frac{1}{3}.$$

Hence, the dc component of the detector output voltage should be determined for effective voltages of

$$u_1 = u_{01} \sin x,$$
$$u_2 = \frac{3}{4} u_{02}\left[\cos(x+\varphi) - \frac{1}{3}\cos 3(x+\varphi)\right], \quad (9)$$

where $u_{02} = [u_2]_{max}$.

A simple fixing of the integration intervals, as in the case of (5), by setting conditions (8) is impossible in this instance, since x_1 and x_2 have to be obtained from equations in which they figure as arguments of functions $\sin^3$ and $\cos^3$.

In this instance, as well as for more complex distortions of voltages u_1 and u_2, the following method is recommended for determining the dc component of the detector output current.

One of the curves, for instance that of u_1, is plotted on millimetric graph paper, and the other (u_2) on tracing paper. The tracing paper is placed over the graph paper in such a manner that their X-axes coincide. Then by displacing the curve of u_2 along the X-axis, definite phase differences between the two voltages are set, the intersection points of the curves are obtained, and according to the rules given in (3) and (4) the outlines of the positive and negative areas which have to be summed up are determined. In Figs. 2 and 4 these areas are shaded. The integration is made by means of a planimeter for various phase angles.

The integration results for (9) are illustrated by curves 3, 4 and 5 in Fig. 3.

Curve 3 corresponds to the case of $u_{01} = \frac{1}{2} u_{02}$, curve 4 to the case of $u_{01} = 2 u_{02}$, and curve 5 to $u_{01} \gg u_{02}$.

It will be seen that in the presence of a third harmonic in voltage u_2 a deviation of angle φ by $\pm 45°$ from its initial value of 90° reduces, under certain conditions, the detector output current only by 7-8% (Curve 5). If the detector is fed with a sinusoidal voltage, the same deviation of angle φ produces a reduction in the current of 30% (curve 1) or even of 45% (curve 2).

SUMMARY

1. For the plotting of graphs of currents in a resistive load of a ring detector fed by arbitrary voltages given in an analytical or graphic form, it is recommended that formulas (3) or (4) be used.

2. When a ring detector is used as an FM demodulator it should be fed with an ac voltage of the same amplitude. The phase characteristic of the detector will then be more linear than if one voltage has a larger amplitude than the other.

3. When the ring detector is used as a phase discriminator for measuring the active component of a bridge output voltage (for instance, in a strain gauge equipment which has long leads to the transducers), the bridge should be fed from a transformer with a saturated core, or else the bridge supply voltage should be distorted by other means with odd harmonics in order to reduce the effect of unstable stray bridge capacitances on the accuracy of measurements.

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MEASUREMENT OF FREQUENCY DEVIATIONS IN FM OSCILLATIONS BY MEANS OF A FREQUENCY METER

P. A. Shpan'on

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A reference method for testing deviations developed by the KhGIMIP (Khar'kov State Institute of Measures and Measuring Instruments) in 1957-59 consists of heterodyning a frequency-modulated signal and measuring by means of a frequency counter the alternating frequency of the signal at the output of a mixer whose mean frequency approaches zero.

The utilization of a frequency counter with a continuous counting over a given frequency interval which is considerably larger than the period of the modulating frequency was suggested by L. D. Bryzzhev in 1957.

For a practical application of this method over a wide range of carrier frequencies it was necessary to examine the relation between the readings of the frequency counter and the measured deviation for any value of the intermediate frequency, and to find the additional measurement error produced by small deviations of the modulating function from a sinusoidal shape.

Only a particular case when the mean intermediate frequency is equal to 0 and the modulation is sinusoidal has been analyzed in print [1] (a capacitative frequency meter was used in this experiment which, however, does not change the quantitative relations).

The present work deals with a more general case.

Theoretical analysis. The FM signal whose deviation is being measured is first heterodyned. It can then be represented in the form:

$$u = U_m \sin\left(\omega_{1f}\, t + \frac{\Delta\omega}{\Omega} \sin \Omega t\right), \qquad (1)$$

where $f_{1f} = \omega_{1f}/2\pi$ is the mean intermediate frequency; $\Delta f = \Delta\omega/2\pi$ is the frequency deviation; $F = \Omega/2\pi$ is the modulation frequency.

Let us find the relation of the meter readings to f_{1f} and Δf when $f_{1f} \leq \Delta f$.

The accumulation of the FM signal phase over a period of the modulating function is determined from expression [2]:

$$\psi = \int_0^T (\omega_{1f} + \Delta\omega \cos \Omega t)\, dt.$$

The number of full phase change periods over this interval will be

$$N' = \frac{\psi}{2\pi} = \mathrm{ent}\left[\frac{1}{2\pi}\int_0^T (\omega_{1f} + \Delta\omega \cos \Omega t)\, dt\right].$$

The symbol ent (integer) is used because the number of times the phase passes through zero can only be expressed by a whole number.

The number of zeros for a rising (or falling) function over a period T and including the first zero of the function is equal to

$$N_T = 1 + \text{ent}\left[\frac{1}{2\pi}\int_0^T (\omega_{1f} + \Delta\omega \cos \Omega t)\, dt\right]. \tag{2}$$

Thus, the readings of the counter over a period of 1 sec will be

$$N = F\left\{1 + \text{ent}\left[\frac{1}{2\pi}\int_0^T (\omega_{1f} + \Delta\omega \cos \Omega t)\, dt\right]\right\}. \tag{3}$$

By separating the integral into three parts we obtain (see figure)

$$N = F\left\{1 + \text{ent}\left[\frac{1}{2\pi}\int_0^a (\omega_{1f} + \Delta\omega \cos \Omega t)\, dt + \frac{1}{2\pi}\int_a^{T-a} (\omega_{1f} + \Delta\omega \cos \Omega t)\, dt + \right.\right.$$

$$\left.\left. + \frac{1}{2\pi}\int_{T-a}^{T} (\omega_{1f} + \Delta\omega \cos \Omega t)\, dt\right]\right\}. \tag{4}$$

The first and third intervals are positive, and the second is negative, which corresponds to the phase accumulation in the interval < a, T-a > proceeding in the reverse direction. Since this does not affect the reading of the instrument, it is necessary to count the modulus of this interval.

The solution of (4) is

$$N = F\left\{1 + \text{ent}\left[\frac{\Delta f}{F\pi/2}\left(\sqrt{1 - \left(\frac{f_{1f}}{\Delta f}\right)^2} + \frac{f_{1f}}{\Delta f} \arcsin \frac{f_{1f}}{\Delta f}\right)\right]\right\}. \tag{5}$$

For condition $\Delta f \gg F$ the error due to the fact that $f_{1f} \neq 0$ will be

$$\delta = \left(\sqrt{1 - \left(\frac{f_{1f}}{\Delta f}\right)^2} + \frac{f_{1f}}{\Delta f} \arcsin \frac{f_{1f}}{\Delta f} - 1\right) 100\,\%. \tag{5a}$$

Let us now examine the limiting cases.

1. Let $f_{1f} = 0$. Then the reading of the frequency counter will be $N = F\left[1 + \text{ent}\left(\frac{\Delta f}{F\pi/2}\right)\right]$. If $\Delta f \leq F$, so that $\frac{f}{F\pi/2} < 1$, then N = F, i.e., the counter does not respond to modulation and its readings are equal to the modulating frequency. If $\Delta f \gg F$ (the case of a slow frequency variation), then $N \approx \frac{\Delta f}{\pi/2}$, since it is possible to neglect unity and the symbol ent with an error of $\frac{\pi}{2(\Delta f/F)}$. This corresponds to the particular case analyzed in [1].

2. Let $f_{1f} = \Delta f$, then $N = F\left[1 + \text{ent}\left(\frac{f_{1f}}{F}\right)\right]$. i.e., the readings of the frequency counter are equal to the mean intermediate frequency (or deviation) for $f_{1f} \gg F$, which is the same as $\Delta f \gg F$.

3. If $f_{1f}/\Delta f$ is small, then (5) can be simplified by expanding it into a series. For $\Delta f / F \gg 1$ we have

$$N = \frac{\Delta f}{\pi/2}\left[1 + \frac{1}{2}\left(\frac{f_{1f}}{\Delta f}\right)^2 + \frac{1}{24}\left(\frac{f_{1f}}{\Delta f}\right)^4 + \ldots\right]. \tag{6}$$

Thus, if during measurement the intermediate frequency has changed by 0.1 of the deviation value, an error of 0.5% will be produced. Since the frequency drift during the measurement time is considerably smaller than the above value, it is possible to consider that the method is sufficiently accurate even with an unstable carrier frequency.

It follows from (3) that for $f_{if} \gg \Delta f$ the counter reading is N = F [1 + ent (f_{if}/F)], i.e, it is equal to the mean FM signal frequency for $f_{if} \gg F$.

This circumstance can be utilized for measuring "up" and "down" deviations in case of asymmetrical modulation for which, in addition to the extreme values of the instantaneous frequency, it is also necessary to measure the mean intermediate frequency.

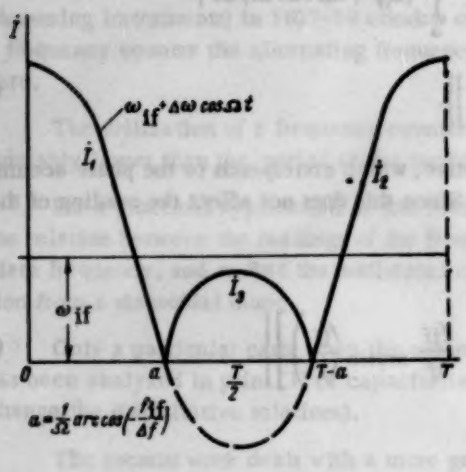
Let us now examine the effect of nonlinear distortions on the frequency counter readings.

Let $f_{if} = 0$ and $\Delta f/F \gg 1$. Let us assume that the frequency varies according to the law:

$$f(t)=\Delta f_1 \sin \Omega t - \Delta f_2 \sin (2\lambda \Omega t+\varphi),$$

where λ is an integer.

We assume the nonlinear distortion coefficient to be small, i.e., $2/\Delta f_1 \ll 1$.



In case the modulation index is large it is possible to consider that in each time interval dt the frequency is constant. Then the counter readings will correspond to the mean frequency, i.e.,

$$N=\frac{1}{T}\int_0^T [\Delta f_1 \sin \Omega t - \Delta f_2 \sin (2\lambda \Omega t+\varphi)]\, dt. \qquad (7)$$

It should be remembered in integrating that the frequency cannot have negative values. Hence, integration is made within the limits for which the integral multiplier is positive, and the result thus obtained is doubled.

The limits can be obtained from an approximate solution of equation

$$\Delta f_1 \sin \Omega t - \Delta f_2 \sin (2\lambda \Omega t+\varphi)=0.$$

Taking the above remarks into consideration, the solution of (7) for $\varphi = \pi/2$ with an accuracy of $(\Delta f_2/\Delta f_1)^2$ amounts to:

$$N=\frac{\Delta f_1}{\pi/2}.$$

It will be seen from this expression that the counter is hardly affected by the second harmonic [the difference in readings with and without the second harmonic is of the order of $(\Delta f_2/\Delta f_1)^2$]. However, an accurate deviation meter would have read in this instance $\Delta f_{up} = \Delta f_1 + \Delta f_2$ and $\Delta f_{dn} = \Delta f_1 - \Delta f_2$, i.e., the measurement error of the above method in the presence of the second harmonic is of the order of the nonlinear distortion coefficient for $\varphi = \pi/2$.

For $\varphi = 0$ the frequency counter does not react to the second harmonic.

Let the frequency vary according to the law

$$f(t)=\Delta f_1 \sin \Omega t+\Delta f_2 \sin (\mu \Omega t+\varphi),$$

where μ is an odd number. In this case the counter reading will be:

$$N=\frac{2\Delta f_1}{\pi}\cos\frac{\Delta f_2 \sin\varphi}{\Delta f_1+\Delta f_2\mu\cos\varphi}+\frac{2\Delta f_2}{\pi\mu}\cos\left(\frac{-\mu\Delta f_2\sin\varphi}{\Delta f_1+\mu\Delta f_2\cos\varphi}+\varphi\right). \qquad (8)$$

For $\varphi = 0$ expression (8) takes the form

$$N=\frac{\Delta f_1+\frac{\Delta f_2}{\mu}}{\pi/2}.$$

In this case an accurate deviation meter would have read $\Delta f_{up} = \Delta f_{dn} = \Delta f_1 - \Delta f_2$ [$\varphi = 0$, worst case].

Thus, the readings of a frequency counter differ from the actual readings by not more than $\Delta f_2 (1-1/\mu)$. For $\varphi = \pi/2$ an accurate deviation meter reads Δf_1, but the frequency counter readings vary by $(\Delta f_2/\Delta f_1)^2$. In this case the measurement error of the frequency meter does not exceed the nonlinear distortion coefficient.

Experimental part. In order to check the absence of any systematic errors in this method it was necessary to make simultaneous measurements by means of this method and some other checking method. The frequency deviation Δf_g was measured with a sinusoidal frequency modulated generator calibrated at various modulation frequencies by means of Bessel function zeros with a random error not exceeding 0.5% and with a frequency counter Ch3-1 after converting the FM signal into an FM signal with an intermediate frequency approaching zero. The measurement results are given in Table 1.

TABLE 1

Δf_g, kg	Meter readings for $\Delta f \approx 0$, kc	$\Delta f = N\pi/2$, kc	Deviation, %	Modulating frequency, cps
20	12.8	20.1	0.5	
25	16.1	25.2	0.8	
30	19.2	30.1	0.3	
35	22.5	35.2	0.6	
40	25.6	40.2	0.5	
45	28.8	45.2	0.4	
50	31.9	50.2	0.4	70
55	35.1	55.0	0.0	
60	38.3	60.3	0.5	
65	41.5	65.1	0.2	
70	44.8	70.4	0.5	
75	48.1	75.3	0.5	

TABLE 2

f_{if}, cps	Error, %	
	calculated from the formula	actual value
1500	0.04	0.0
5500	0.5	0.3
10000	1.5	1.1
12000	2.3	2.0
15000	2.6	3.0
20000	7.6	6.7

It will be seen from the data given in Table 1 that the measurement results obtained by the two methods do not differ by more than 0.8%.

In order to check the accuracy of (5) a set deviation value (55 kc) was measured at various intermediate frequencies. The measurement results are given in Table 2.

The data shown in Table 2 confirm that the additional error in deviation measurements due to the drift in the intermediate frequency during the measurement is very small.

SUMMARY

It follows from the theoretical and experimental investigations that the error of measurement by the above method for a zero intermediate frequency and a sinusoidal modulation is equal to $\frac{\pi}{2(\Delta f/F)} \cdot 100\%$, i.e., it drops with a rising modulation index. The deviation of the intermediate frequency from zero has a small effect on the measurement results [the additional error due to this factor is of the order of $(f_{if}/\Delta f)^2$]. Since the error of the

frequency counter can normally be neglected, the only source of error for a large modulation index, in fact, consists of the modulating function deviation from a sinusoidal shape. For a small nonlinear distortion coefficient this error is of the same order as the coefficient.

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METHOD FOR MEASURING FREQUENCY DEVIATION

V. L. Rychka

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Several methods have been described [1] for measuring the frequency deviation of frequency-modulated signals; however, their measurement errors exceed 2-5%, i.e., they exceed the value specified for checking FM ultrashortwave frequency deviation meters and standard signal generator modulation meters.

The problem is complicated by the fact that it is often necessary to measure the frequency errors of a frequency deviation meter over a wide range of modulating frequencies. For instance, the frequency characteristic of a modern USW FM broadcasting transmitter must be straight over a range of 30-15,000 cps with an error of ± 1 db (± 6%). Even more stringent requirements are specified for measuring receivers and the means for checking them.

These reasons led to the necessity of improving the technique of certain known methods so as to attain the required accuracy.

In this article we examine the method of zeros in the FM signal amplitude characteristic based on the utilization of the Bessel function zeros.

In order to raise the accuracy of the readout, increase the number of readout points and eliminate other defects of this method, certain alterations have been introduced which are described below.

Description of the method. The method is based on the disappearance of certain frequencies in the FM signal spectrum for certain values of the modulation index. This is a well-known technique and will not be described here. It is also known that this method has its defects, namely a relatively small number of readout points, a limiting minimum value of the modulation index, its applicability only to the upper modulation frequencies, and finally, its low accuracy of the order of 10%. It has been pointed out in literature that the latter fact is due to its insufficiently reliable readings.

Let us first of all note that the number of readout points can be increased considerably by using Bessel function zeros not only of the zero order, but also of the first and second order. It is known from the theory of the FM frequency characteristics that the Bessel function of the zero order $J_0(\beta)$ (where β is the modulation index) determines the amplitude of the center frequency oscillations, the Bessel function of the first order $J_1(\beta)$ determines the amplitude of the first sideband frequency, that of the second order $J_2(\beta)$ the second sideband, etc. The zeros of these functions are interspaced for larger modulation indexes. Hence, if we observe the amplitude zeros of these oscillations we can obtain a large number of reference points sufficient for checking various scales of the frequency deviation meters and for plotting frequency characteristics.

For convenience of determining the actual maximum frequency deviation for various modulating frequencies the table shows calculation results based on the Bessel function roots of the zero, first and second order. The values of the maximum deviation are obtained with an error not exceeding 0.5%. If still greater accuracy is required one of the Bessel function tables should be used.

The table only provides the first few roots of the second order Bessel functions, since with a rising number of the root its value approaches the corresponding root of the zero order Bessel function. For the same reasons there was no point in considering roots of Bessel function of a higher order.

The table shows the frequency deviations for modulating frequencies of 1 to 7 kc in steps of 1 kc. When it is required to calculate deviations for intermediate frequencies, figures in the fourth column of the table (deviation for a modulating frequency of 1 kc) should be multiplied by the required frequency in kilocycles.

Actual Frequency Deviations for Zero Values of the Bessel Function

Bessel function order	Oscillation frequency	Root No.	Modulating frequency, kc														
			1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
0	Center	1	2.4	4.8	7.2	9.6	12.0	14.4	16.8	19.2	21.6	24.0	26.4	28.8	31 2	33.6	36
1	First sideband	1	3.83	7.66	11.5	15.3	19,1	23.0	26,8	30,6	34,5	38.3	42,1	46.0	49.8	53.6	57.5
2	Sec. sideband	1	5.15	10.3	15.5	20.6	25,8	30.9	36.0	41.2	46.4	51.5	56,6	61.8	67,0	72,1	
0	Center	2	5.52	11.04	16.6	22.1	27,6	33,1	38.6	44.2	49.7	55,2	60.7	66,2	71.8	77.3	
1	First sideband	2	7.0	14.0	21.0	28.0	35,0	42.1	49,1	56.1	63,1	70,1	77 1				
2	Sec. sideband	2	8.42	16.8	25.3	33.7	42,1	50.5	58.9	67.4	75,8	84,2					
0	Center	3	8.65	17.3	26.0	34.6	43,2	51.9	60.6	69.2	77.9						
1	First sideband	3	10.2	20.4	30.6	40.8	51,0	61.2	71.4	81.6							
0	Center	4	11.8	23.6	35.4	47.2	59,0	70.8	82.6								
1	First sideband	4	13.3	26.6	40.0	53.2	66,5	79.8									
0	Center	5	14.9	29.8	44.7	59.6	74,5										
1	First sideband	5	16.5	32.9	49.3	65.8											
0	Center	6	18.1	36.2	54.3	72.4											
1	First sideband	6	19.6	39.2	58.8	78.4											
0	Center	7	21.2	42.4	63.6												

The table only includes modulating frequencies not less than 1 kc. The application of this method to modulating frequencies below 1 kc is difficult, mainly owing to the fact that the side frequencies spectrum becomes very congested and the tuning of the receiver to a given side frequency becomes difficult owing to the instability of the FM generator and the local heterodyne oscillator. Even at a modulating frequency of 1 kc an especially careful measurement technique is required (including preheating of the measuring apparatus) in order to avoid errors in determining the number of the side frequency to which the receiver is tuned. For a modulating frequency of 1 kc and higher it is possible to tune to 7-8 side frequencies. The measurements are made considerably easier if the source of FM oscillation consists of a highly stable, for instance, crystal-quartz, generator.

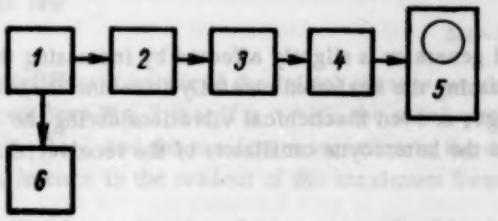

Fig. 1. 1) Audio oscillator; 2) ultra-shortwave generator type GSS; 3) double superheterodyne radio receiver; 4) low-frequency band filter; 5) cathode-ray oscilloscope type ÉO-7; 6) quartz-crystal calibrator.

Measurement technique. For checking deviation meters an FM signal is fed simultaneously to the instruments under test and a reference frequency deviation meter which locates the Bessel function zeros. Such reference instruments described in literature consist of superheterodyne receivers with intermediate frequency quartz filters and an output audio indicator.

In order to raise the accuracy of the readout and make the operation more convenient, a modified circuit with an oscilloscope indicator was used.

The reference instrument consisted of a normal superheterodyne wide-band receiver with the required frequency range and an adjustable beat oscillator used to obtain zero beats between the intermediate frequency and the second heterodyne. The beat oscillator should be tunable over a frequency range equal to at least double the maximum modulating frequency.

Such a receiver may consist, for instance, of an ultra-shortwave frequency deviation meter type 185. The block schematic of this instrument is shown in Fig. 1.

The beat frequency voltage is fed from the receiver through a narrow-band low-frequency filter to the input of a cathode-ray oscilloscope. The filter is required to eliminate the modulating frequency and pick out the beat

frequency voltage. For this purpose it is only necessary to have a low-pass filter with a cut-off frequency below 1 kc and large attenuation outside the passband. However, with a narrow-band filter it is more convenient to observe the beating on the oscilloscope screen.

For this purpose an RC band filter, for instance, can be used. The width of the passband and flatness of its characteristic are not important; however, a very sharply tuned filter (with a bandwidth less than a few dozen cps) is not convenient for operation, since the slightest variations in the frequency of the ultra-shortwave generator, or that of the first and second heterodyne oscillators in the receiver produce a considerable voltage drop at the filter output. The filter transfer constant should not be less than 0.1, otherwise the receiver output voltage of 1-2 v would not be sufficient for the existing sensitivity of the oscilloscope to produce a sufficiently accurate determination of the Bessel function roots.

Experience has shown that it is best to use mains synchronization for the oscilloscope.

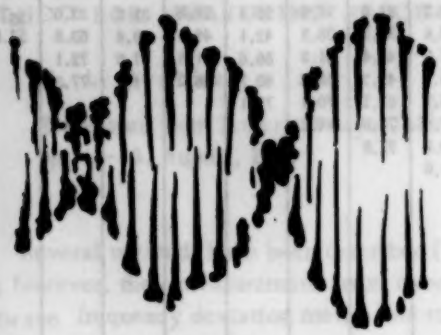

Fig. 2.

When the modulation index attains the value of a Bessel function root the oscilloscope image should appear as a horizontal line. However, for the same reason as the modulation index changes slightly with the mains frequency (or double that frequency), the oscilloscope image takes the shape shown in Fig. 2 due to the pulsations of the supply voltage.

At the instant the modulation index is equal to the Bessel function root, the amplitude of oscillations becomes equal to zero.

The measurement technique consists of the following. Having switched off the generator modulation we tune in the receiver and set the frequency of the beat oscillator so as to obtain a maximum signal amplitude on the oscilloscope. If a tuned filter is used we shall obtain two maxima corresponding to the tuning on either side of the zero beats. Either of these positions can be used. The generator modulation is then switched in and the deviation smoothly increased from zero. If the actual value of deviation which it is required to find corresponds to a root of the zero order Bessel function, the tuning of the beat frequency oscillator should not be changed, but by increasing the deviation the number of the root should be counted by observing the disappearance of the beat frequencies on the oscilloscope screen.

When, however, the deviation corresponds to a Bessel function root of the first or second order, it is advisable after tuning in the receiver and the beat oscillator first to set a small deviation sufficient for observing the side frequencies, but small enough not to miss any Bessel function zeros. The beat frequency oscillator is then returned in such a manner as to obtain a maximum output voltage for the required side frequency, and then measurements are carried out in the same manner as in the preceding case.

It should be noted that the center frequency of the USW signal generator is slightly affected by increasing the frequency deviation; hence, it is necessary simultaneously with increasing the deviation carefully to retune the beat frequency oscillator. Moreover, small variations of the supply voltage, or even mechanical vibrations during the experiment may change eventually the frequency of the generator or the heterodyne oscillators of the receiver, thus requiring retuning of the beat frequency oscillator.

The accuracy in setting a given deviation frequency is determined by the accuracy of setting the modulation frequency, the accuracy of observing the disappearance of beats, and the accuracy with which the corresponding Bessel function root is calculated.

In order to raise the accuracy of setting the modulation frequency it is advisable to use a crystal-quartz oscillator type 596-A. Since the actual deviation is equal to the product of the modulating frequency and the modulation index which is the root of the function, the deviation error will equal the sum of the errors of these two quantities. The modulation frequency can be set with an error of 0.1%. The Bessel function root can be calculated with any accuracy. The table provides numerical values of the roots with an error not exceeding 0.5%. Hence, the error in determining the frequency deviation by means of the method shown in the block schematic (Fig. 1) is almost entirely determined by the error with which the disappearance of beats can be observed on the oscilloscope screen. The greater the amplification of the beat voltages and the higher the stability of the frequency generator and the heterodyne oscillators, the greater will be the accuracy. Experiments carried out with the equipment shown in the block

schematic (Fig. 1) show that, after a preliminary heating up period, it is possible to observe with certainty, for a variation of the modulation frequency by 1%, the disappearance or reappearance of beat voltages up to the 7-8th root of any function. Hence, the total error in setting the frequency deviation by means of this method does not exceed 1-2%, which can be considered amply sufficient for checking frequency deviation meters at modulating frequencies above 1 kc. The errors of this method are examined in greater detail in [2].

Effect of nonlinear distortions. Let us now examine the effect of nonlinear distortions in the FM signal on the measurements of the frequency deviation by means of this method, which is based on bringing to zero certain components of the spectrum.

Let the instantaneous frequency of the FM signal follow the law

$$\omega(t)=\omega_0+\Delta\omega_1\sin\Omega t+\Delta\omega_2\sin 2\Omega t.$$

The complete phase of oscillations in time t will be:

$$\varphi(t)=\int_0^t\omega(t)dt=\omega_0 t-\beta_1\cos\Omega t-\beta_2\cos 2\Omega t.$$

It is known [3] that for a simultaneous modulation with two frequencies the FM frequency spectrum contains the following components:

a) the carrier frequency ω_0 with an amplitude proportional to the product of zero order Bessel functions with arguments β_1 and β_2;

b) the side frequency $\omega_0 \pm n\Omega$ with amplitudes proportional to the product of $J_n(\beta_1) \cdot J_0(\beta_2)$;

c) additional side frequencies of the form $\omega_0 \pm (p\Omega + 2n\Omega)$.

It should be noted that if the nonlinear distortions do not exceed a few percent argument $\beta_2 < 2.4$ even for $\beta_1 = 50$. This means that the second factor in the above products virtually never becomes zero for modulating frequencies above 1 kc.

For this reason small nonlinear distortions do not provide additional zeros in the side frequency component amplitudes with a varying modulation index.

Nonlinear distortions cause certain errors in measuring the deviation frequency, which, however, do not exceed according to [2] the value of the nonlinear distortion coefficient.

In order to evaluate this error let us examine, for instance, the case when the frequency varies according to the law

$$\Delta\omega=\Delta\omega_1\sin\Omega t+\Delta\omega_2\sin(2\Omega t+\varphi).$$

For different values of the initial phase φ we shall obtain an asymmetrical frequency modulation. This, it will be seen from Fig. 3 that if $\varphi = \pi/2$, the maximum deviation frequency in the direction of rising frequencies will equal $\Delta\omega_1-\Delta\omega_2$, and the maximum variation in the direction of falling frequencies will equal $\Delta\omega_1 + \Delta\omega_2$. Hence, the difference in the readout of the maximum frequency deviation is equal to $2\Delta\omega_2$.

Fig. 3. Generator frequency variations in the presence of nonlinear distortions.

The above method is intended mainly for checking deviation meters. Since deviation meters usually measure the peak value of the deviation, their error due to nonlinear distortions will also be of the order to nonlinear distortion coefficient. When, for instance, generator type GSS-17 is used, whose nonlinear distortions do not exceed 1%, the measurement error due to nonlinear distortions does not exceed 2-3%.

If generators GSS-17 or GSS-30 are used, the following most important fact must be borne in mind. The generator inputs should be fed with an external modulation from an audio oscillator. However, the peculiarity of the generator circuits consist in the fact that during external modulation the internal source of the sinusoidal voltage is not switched off; hence, owing to stray couplings the external modulating

signal is superposed by a small internal signal, thus considerably impeding the observation of Bessel function zeros. Normally it is advisable to open the generator and remove the internal audio oscillator tube.

Similar difficulties arise when the FM signal generator has a considerable so-called FM background, i.e., a stray frequency modulation by means of the rectifier background, or by mains frequency pickup at the input of the reactance tube. This background must be at least 40 db below the nominal deviation level, which is normally easily attainable in commercial generators GSS-17 and GSS-30.

It has been noted above that this method is only suitable for measuring deviations with modulating frequencies exceeding 1-2 kc. For measuring deviations with lower modulating frequencies with the same accuracy we have developed an appropriate method which will be described in due time.

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DIRECT READING INSTRUMENT FOR MEASURING LOSS-ANGLE TANGENT IN CAPACITORS

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The mass production of capacitors requires mechanization and automation for checking their parameters. The measurement of tg δ normally requires instruments with tuning facilities (bridges, tuned circuits), and involves calculations, thus increasing the time spent on the measurements. These defects have been eliminated in instruments IPP-1, IPP-2 and IPP-3 [1, 2, 3]. The analysis of the theory underlying these instruments is therefore of interest, since it will facilitate and improve the designing of similar devices.

The simplified schematic of these instruments (Fig. 1) consists of a dc bridge made up of resistances R_1, R_2, and R_3, and the dc resistance of the high-frequency oscillator tube T. When the tested capacitor C_X is connected to the oscillator tuned circuit at terminals ab the bridge becomes unbalanced. Before connecting capacitor C_X, capacitor C_V is set in a position corresponding to the measured capacitance C_x in such a manner that the total capacitance of the tuned circuit C_k remains constant for any value of C_X. Capacitor C_V has a scale indicating its position in measuring any value of C_X. Capacitance C_X must be known in advance. The instrument is designed for measuring tg δ of capacitors which have already been tested for capacitance; hence, the above condition does not lower the operational qualities of the instrument.

The current which flows in the bridge diagonal when the tested capacitor C_X is connected is proportional to tg δ. This is attained by an adjustment of the diagonal resistance which corresponds to capacitance C_X, and is obtained automatically since resistor R_0 is ganged to capacitor C_V.

In order to obtain proportionality between tg δ and the current in the diagonal, the circuit parameters must satisfy certain conditions which will be specified below.

Let the rotation angle γ of capacitor C_V vary in the range of 0 to 1. Moreover, let the total capacitance C_k of the oscillator tuned circuit be expressed by equation

$$C_k = C_n(1-\gamma) + C_p \frac{\alpha}{\alpha+\beta}, \qquad (1)$$

where C_n is the maximum value of capacitor C_v; C_p is the capacitance connected in series with the tested capacitor; C_{xn} is the largest (nominal) value of the capacitance whose tg δ can be measured by means of this set; $\beta = C_p/C_{xn}$; $\alpha = C_x/C_{xn}$.

Considering that the total tuned circuit capacitance C_k is constant for any values of γ, we can obtain after tedious transformations, which have been omitted here, the following equation:

$$C_n = C_p\left(\frac{1}{1+\beta} - \frac{\alpha_{min}}{\alpha_{min}+\beta}\right), \qquad (2)$$

$$\alpha = \frac{\alpha_{min}(1+\beta)+\beta\gamma(1-\alpha_{min})}{1+\beta-\gamma(1-\alpha_{min})}, \qquad (3)$$

where α_{min} is the value of α for $\gamma = 0$.

The graph of the relationship between α and γ as represented by (3) is shown in Fig. 2, with α_{min} equal to 0.1 and 0.3 denoted by full and broken lines, respectively.

For a relative variation of conductance equal to $\Delta g/g$, current I in the bridge diagonal is equal to [4]:

$$I = \frac{I_{a0} R_a k'_p \frac{\Delta g}{g}}{R + \frac{R_a R_l}{R_a+R_l} + \frac{R_1 R_2}{R_1+R_2}}, \qquad (4)$$

where I_{a0} is the dc component of the oscillator anode current; R is the bridge diagonal resistance; $k'_p = \frac{I_{a0}}{I_{a0}} \frac{g}{\Delta g}$ is a coefficient which depends on the oscillator's operating condition.

Coefficient k'_p relates the oscillator anode current relative variation $\Delta I_{a0}/I_{a0}$ to $\Delta g/g$. The value of k'_p is determined from the load characteristic of the oscillator. For an overloaded condition, k'_p approaches unity.

Let us now express the increment of the tuned circuit conductance Δg in terms of tg δ of the tested capacitor, namely tg $\delta = r_x \omega C_x$, where r_x is a series resistance equivalent to the losses in capacitor C_x. For this purpose let us convert the series equivalent circuit to the right of points cd (Fig. 1) into a parallel circuit (Fig. 1a). For these circuits the loss angles must be the same.

For small values of δ we have

$$r_x \omega \frac{C_p C_x}{C_p + C_x} = \frac{\Delta g}{\omega \frac{C_p C_x}{C_p + C_x}}.$$

Hence,

$$\Delta g = \operatorname{tg}\delta \frac{\omega C_p}{k}, \qquad (5)$$

where

$$k = \frac{(\beta+\alpha)^2}{\beta\alpha}.$$

From (3) we obtain

$$k = \frac{1}{D + F\gamma - H\gamma^2},$$

where

$$D = \frac{\beta\alpha_{min}}{(\beta+\alpha_{min})^2}; \quad F = \frac{\beta(1-\alpha_{min})(\beta-\alpha_{min})}{(1+\beta)(\beta+\alpha_{min})^2}; \quad H = \frac{\beta^2(1-\alpha_{min})^2}{(\beta+\alpha_{min})^2(1+\beta)^2}.$$

By inserting (5) into (4) we obtain

$$I=\operatorname{tg}\delta\,\frac{I_{a0}R_a\,k_p'\omega C_p}{gkN}, \tag{7}$$

where

$$N=R+\frac{R_aR_l}{R_a+R_l}+\frac{R_1R_2}{R_1+R_2}.$$

In order to calibrate meter M in values of tg δ it is necessary to have a constant coefficient of tg δ in (7) for any value of C_x. All the parameters in the coefficient of tg δ with the exception of kN are not affected by C_x; it is possible to select the circuit parameters in such a manner that the product kN is independent of C_x. Let us find the conditions for which this holds.

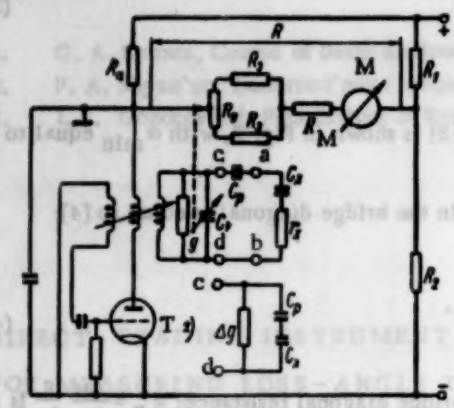

Fig. 1.

It is easy to express the diagonal resistance R as a function of γ. We shall then have

$$N=A+B(\gamma-\gamma^2), \tag{8}$$

where

$$A=\frac{R_3^2+R_3R_0}{R_0+2R_3}+R_U+\frac{R_aR_l}{R_a+R_l}+\frac{R_1R_2}{R_1+R_2}; \tag{9}$$

$$B=\frac{R_0^2}{R_0+2R_3}. \tag{10}$$

It is easy to discover by examining equations (6) and (8) that the product kN will not depend on γ and will be equal to

$$Nk=\frac{A}{D}=C=\text{const}, \tag{11}$$

if the following conditions are fulfilled:

$$F=H;\quad \frac{A}{B}=\frac{F}{D}. \tag{12}$$

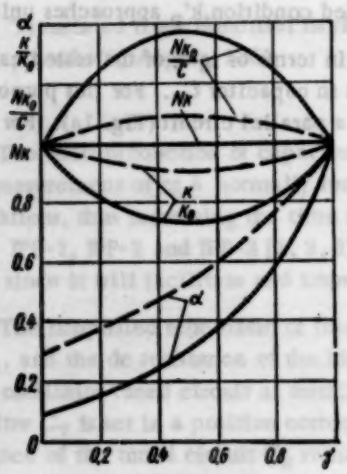

Fig. 2.

By inserting the values of D, F, and H into (12) we shall find the condition for which the readings of meter M will be proportional to tg δ for any value of C_x

$$\beta=\sqrt{\alpha_{\min}}, \tag{13}$$

$$\frac{B}{A}=\frac{(1-\sqrt{\alpha_{\min}})^2}{\sqrt{\alpha_{\min}}}. \tag{14}$$

The above relations make it possible to express in terms of $\alpha_{\min}$ basic functions which relate to circuit parameters, thus making the computation of the circuit considerably easier in designing the instrument.

Taking into consideration (12) and (13) we find from (6)

$$\frac{k}{k_0}=\frac{1}{1+\dfrac{(1-\sqrt{\alpha_{\min}})^2}{\sqrt{\alpha_{\min}}}(\gamma-\gamma^2)}, \tag{15}$$

where k_0 is the value of k for γ = 0 (or γ = 1).

From (1) and (15) we have

$$\frac{Nk_0}{C} = 1 + \frac{(1-\sqrt{\alpha_{min}})^2}{\sqrt{\alpha_{min}}}(\gamma-\gamma^2). \tag{16}$$

The graphs of relations (15) and (16) are shown in Fig. 2 for $\alpha_{min} = 0.1$ (full line) and for $\alpha_{min} = 0.3$ (dotted line).

From (2) and (13) we find

$$C_p = \sqrt{\alpha_{min}} C_{xn}.$$

Taking into consideration (13) we can simplify expression (2) in the following manner

$$C_x = C_p \frac{1-\sqrt{\alpha_{min}}}{1+\sqrt{\alpha_{min}}}.$$

Taking into consideration (13) and assuming in (1) that $\gamma = 1$ and $\alpha = 1$, we find

$$C_x = C_p \frac{1}{1+\sqrt{\alpha_{min}}}. \tag{17}$$

Equation (7) can be reduced to a more convenient form for practical application if the values of $\underline{k}$ and N obtained from (6) and (8) respectively for any value of γ, for instance $\gamma = 1$, are inserted and relations (13) and (17) are utilized:

$$1 - \text{tg}\,\delta k_p' Q l_{a0} R_a \frac{\sqrt{\alpha_{min}}}{A(1+\sqrt{\alpha_{min}})}, \tag{18}$$

where $Q = \omega C_k / g$ is the Q-factor of the oscillator tuned circuit.

The above analysis makes it possible to pass certain remarks on a rational method for designing this instrument.

At present relationship (14) is obtained by selecting an appropriate value for resistor R_g. It will be seen from (9) and (10) that R_g affects both A and B, thus making its selection difficult. It seems preferable to make $R_g = 0$ and assuming the value of R_0 obtain a certain value for B. Then, by connecting a resistance into the diagonal in series with R_M, a value for A can be obtained, which satisfies relation (14).

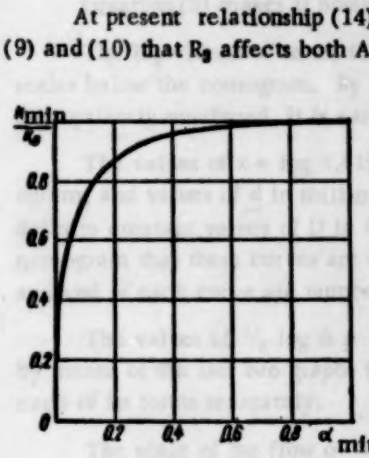

Fig. 3.

The circuit parameters should be chosen in such a way that the adjustment of the sensitivity of meter M and the bridge balancing should not affect the value of A during testing.

It will be seen from the graph in Fig. 2 that a reduction in the capacitance measuring range, i.e., an increase in $\alpha_{min} = C_{x\,min}/C_{nx}$ leads to a reduction of the effect of γ on k/k_0. Hence, it will be less important to compensate the readings by means of R_0. This will lead to a reduction in the compensation error, thus increasing the accuracy of measurement. The minimum value for k/k_0 = k_{min}/k_0 occurs when $\gamma = 0.5$, we shall then have according to (15):

$$\frac{k_{min}}{k_0} = \frac{1}{1+0.25\frac{(1-\sqrt{\alpha_{min}})^2}{\sqrt{\alpha_{min}}}}$$

The above relationship is shown in Fig. 3 from which it will be seen that with a rising α_{min} the value of k_{min}/k_0 approaches unity. It is possible that in certain instances it is advisable to increase α_{min} in order to raise the accuracy of measurements.

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GRAPHIC METHODS FOR COMPUTING CERTAIN PARAMETERS OF FLOWMETERS WITH NORMAL DIAPHRAGMS

R. Z. Alimov

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Graphic methods of computing certain flowmeter parameters can save a considerable amount of time and simplify calculations and checking. These methods are especially valuable for preliminary tentative evaluation of the basic parameters, whose determination often entails fairly complicated calculations.

This paper aims at describing two graphical methods which are used in practice, but are little known to wide circles of experts.

The first method consists of the following.*

The equation for differential manometer flowmeters operating with normal diaphragms is [2],

$$G=0.01252\varepsilon\alpha m D^2\sqrt{\Delta p\gamma}. \quad (1)$$

and can be written in the form:

$$\frac{G}{\varepsilon}\cdot\frac{1}{\gamma^{1/2}}=0.01252\alpha m D^2\Delta p^{1/2}. \quad (2)$$

In a logarithmic form expression (2) becomes:

$$0.5\log\frac{1}{\gamma}+\log G/\varepsilon=\log\ 0.01252\alpha m D^2+0.5\log\Delta p. \quad (3)$$

Equation (3) makes it possible to plot a nomogram (Fig. 1) in the following manner.

The logarithms of terms in the left-hand side of equation (3), i.e., 0.5 log $1/\gamma$ and log G/ε are plotted on scales below the nomogram. By means of these scales, which are placed one after the other in a straight line and appropriately numbered, it is easy to determine the sum of the above terms, that is the left-hand side of equation(3).

The values of $x = \log 0.01252\ \alpha m D^2$ are plotted along the horizontal axis of the right-hand side of the nomogram, and values of $\underline{d}$ in millimeters are plotted along the vertical axis. Curves relating $\underline{d}$ to $\underline{x}$ are plotted for definite constant values of D in the range of 100 to 1000 mm in steps of 50 or 100 mm. It will be seen from the nomogram that these curves are completely identical and only differ by a parallel displacement. The beginning and end of each curve are numbered with the value of D in mm.

The values of $\frac{1}{2}\log\Delta p$ are plotted along the horizontal axis of the left-hand side of the nomogram. Thus, by means of the last two graphs it is possible to determine the value of the right-hand side of equation (3) and of each of its terms separately.

The scale of the flow or G/ε has three ranges, the upper, middle and lower ranges. Numerical values of $\underline{d}$ and D are only indicated in the nomogram for the middle range. When the G/ε values referring to the upper range are used, the values of $\underline{d}$ and D indicated in the graphs should be multiplied by 10, and when the lower range is used they should be divided by 10. The values of γ, V and p are effective for all the flow-measuring ranges. This circumstance is due to the quadratic relationship of G to $\underline{d}$ and D.

Single parameters are determined by means of the nomogram in the following manner.

* A detailed description of this method is given in [1].

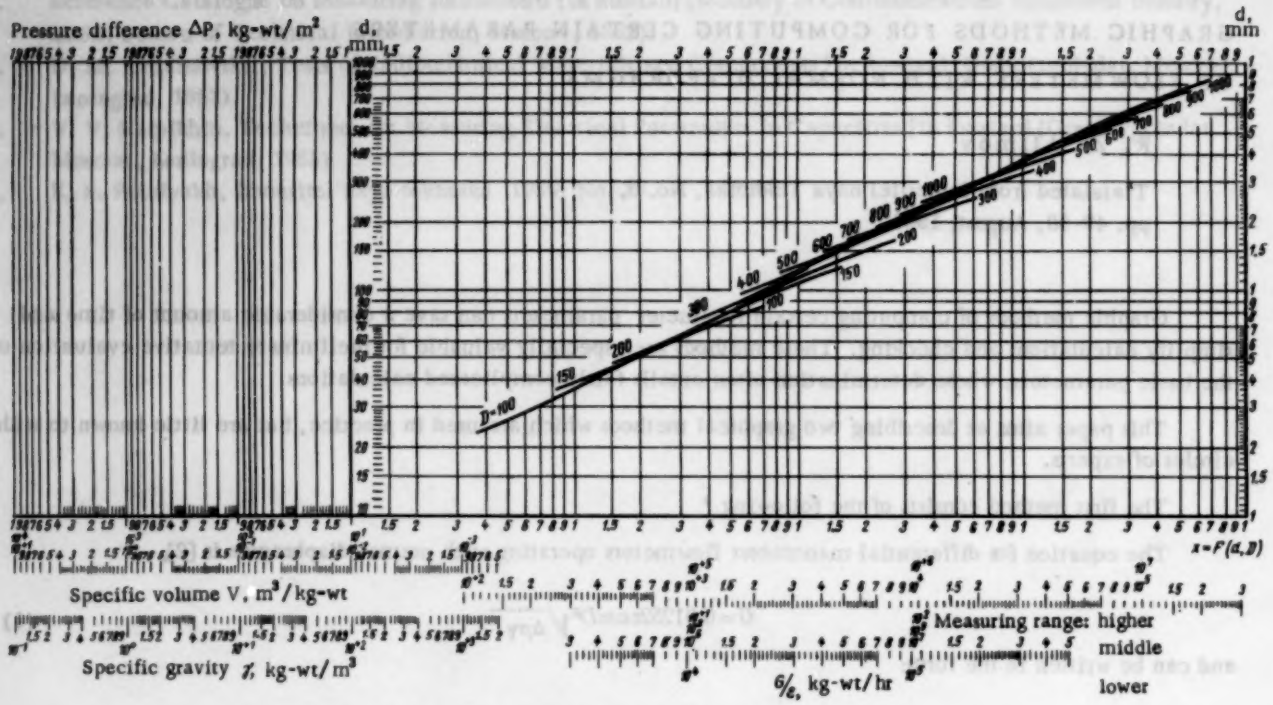

Fig. 1. Nomogram for tentative computations of diaphragms.

For determining d the left end of a transparent paper strip is placed against the given value of γ or V, and on the right-hand side the strip is marked in the placed corresponding to the given value of G/ε (assuming that $\varepsilon = 1$). Thus the sum of $0.5 \log 1/\gamma + \log G/\varepsilon$ is marked off on the strip. The left end of the strip is then placed against the given value of Δp, and in this position the strip is moved upwards parallel to the horizontal axis until the right-hand mark intersects the curve with the given diameter D. The point of intersection with the vertical axis will then indicate the required value of d. It can easily be ascertained that for this value of d the sum $^1/_2 \log 1/\gamma + \log G/\varepsilon$ becomes equal to $^1/_2 \log \Delta p + \log x$, i.e., the graphically determined value of d satisfies Eq. (1) for given values of flow G, specific gravity of the measured medium γ, internal diameter of pipeline D, and the difference of pressures in the constricting device Δp. By determining the value of ε in the normal manner this computation can be made more accurate.

In a similar manner the value of G/ε can be determined for given values of Δp, d, D and γ. For this purpose the left end of the paper strip should be placed against the given value of Δp, the strip located horizontally at the level of the given value of d, and a mark should be made at the point of intersection of the strip with the curve of the given value of D. Thus, a length equal to $^1/_2 \log \Delta p + \log x$ will be marked off on the strip. It has been shown that this length is equal to the sum of $^1/_2 \log 1/\gamma + \log G/\varepsilon$. If the left end of the strip is now placed against the given value of γ on the lower scale, the mark on the strip will indicate on the flow seale the value of G/ε.

There is no difficulty in finding the values of Δp or D from given values of the remaining parameters.

In the second method the graph shown in Fig. 2 is used, which consists of a nomogram for determining the upper limit in the pressure difference of differential manometers, and the loss of pressure in the constricting device with respect to parameter $C = \alpha m \cdot \sqrt{\Delta p}$. By means of this nomogram it becomes possible to evaluate the flowmeter basic parameters for a limited pressure difference in the constricting device.*

The nomogram is plotted from the flow equation and the relation between the loss of pressure P_n and m which is graphically shown in Fig. 3 [2]. The value ($m\alpha\sqrt{\Delta p}$) denoted by C and consisting of the product of three param-

*In the examples only those physical parameters are given which are required for computations by means of nomograms.

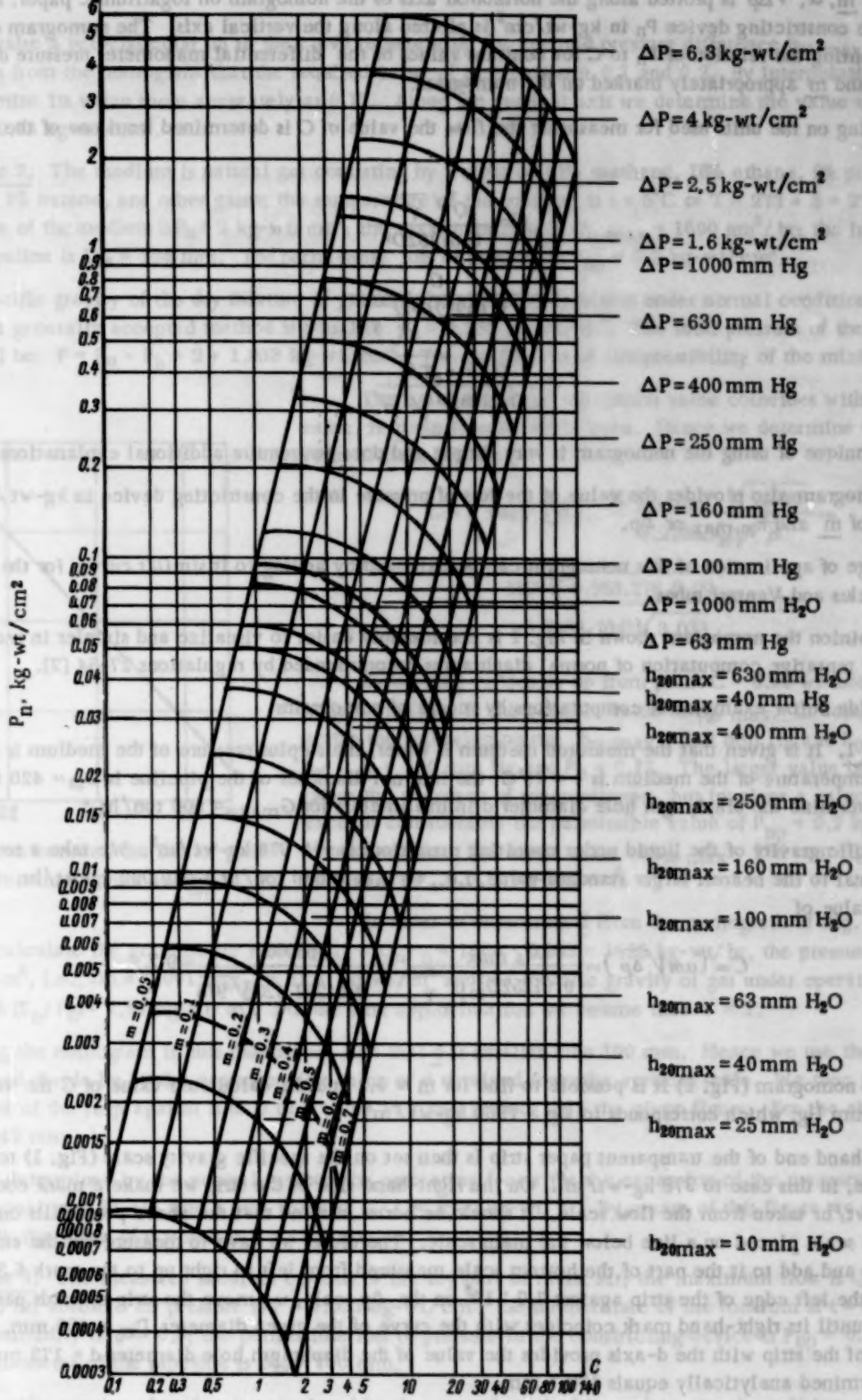
ΔP=6.3 kg-wt/cm²
ΔP=4 kg-wt/cm²
ΔP=2.5 kg-wt/cm²
ΔP=1.6 kg-wt/cm²
ΔP=1000 mm Hg
ΔP=630 mm Hg
ΔP=400 mm Hg
ΔP=250 mm Hg
ΔP=160 mm Hg
ΔP=100 mm Hg
ΔP=1000 mm H₂O
ΔP=63 mm Hg
h20max=630 mm H₂O
h20max=40 mm Hg
h20max=400 mm H₂O
h20max=250 mm H₂O
h20max=160 mm H₂O
h20max=100 mm H₂O
h20max=63 mm H₂O
h20max=40 mm H₂O
h20max=25 mm H₂O
h20max=10 mm H₂O
B=0.05
B=0.1
B=0.2
B=0.3
B=0.4
B=0.5
B=0.6
B=0.7
Pn, kg-wt/cm²
C

Fig. 2.

eters, namely $\underline{m}$, α, $\sqrt{\Delta p}$ is plotted along the horizontal axis of the nomogram on logarithmic paper, and the loss of pressure in the constricting device P_n in kg-wt/cm^2 is plotted along the vertical axis. The nomogram contains curves representing the relation of P_n to C for constant values of the differential manometer pressure differences h_{20} max. Δp and $\underline{m}$ appropriately marked on the nomogram.

Depending on the units used for measuring the flow the value of C is determined from one of the following relations:

$$C=\frac{Q\sqrt{\gamma}}{0.01252D^2}. \tag{4}$$

$$C=\frac{G}{0.01252D^2\sqrt{\gamma}} \tag{5}$$

$$C=\frac{Q_n\sqrt{\gamma_n TK}}{0.2108D^2\sqrt{P}}. \tag{6}$$

The technique of using the nomogram is very simple and does not require additional explanations.

The nomogram also provides the value of the loss of pressure in the constricting device in kg-wt/cm^2 for known values of $\underline{m}$ and h_{20} max or Δp.

The range of application of this nomogram can be extended by adding to it similar curves for the computation of normal nozzles and Venturi tubes.

In our opinion the nomogram shown in Fig. 1 is considerably easier to visualize and simpler in use than the nomograms for tentative computation of normal diaphragms recommended by regulations 27-54 [2].

We provide below examples of computations by means of nomograms.

Example 1. It is given that the measured medium is water, the surplus pressure of the medium is P_u = 4 kg-wt/cm^2, the temperature of the medium is t = 70°C, the internal diameter of the pipeline is D_{20} = 420 mm. It is required to determine the diaphragm hole diameter $\underline{d}$ in millimeters for G_{max} = 600 ton/hr.*

The specific gravity of the liquid under operating conditions is γ = 978 kg-wt/m^3. We take a round figure for the flow equal to the nearest larger standard value, i.e., $G_{s\,max}$ = 630 ton/hr = 630,000 kg-wt/hr. From (5) we calculate the value of

$$C=(\alpha m\sqrt{\Delta p})=\frac{G_{s\,max}}{0.01252D_{20}^2\sqrt{\gamma}}=\frac{630000}{0.01252\cdot 420\sqrt{978}}=9.2.$$

From the nomogram (Fig. 2) it is possible to find for m ≈ 0.2 and the calculated value of C the value of $h_{20\,max}$ = 630 mm Hg, which corresponds to Δp = 7900 kg-wt/m^2.

The left-hand end of the transparent paper strip is then set on the specific gravity scale (Fig. 1) to the appropriate value, i.e, in this case to 978 kg-wt/m^3. On the right-hand side of the strip we make a mark corresponding to 630,000 kg-wt/hr taken from the flow scale. It should be borne in mind that the above point falls on the continuation of the scale placed on a line below the main scale. Therefore, we have to measure out the entire length of the top scale and add to it the part of the bottom scale measured from left to right up to the mark $6.3\cdot 10^5$. Having placed the left edge of the strip against $7.9\cdot 10^3$ on the Δp scale, we move the strip upwards parallel to the horizontal axis until its right-hand mark coincides with the curve of the given diameter D_{20} ≈ 400 mm. The point of intersection of the strip with the d-axis provides the value of the diaphragm hole diameter d = 173 mm. The value of $\underline{d}$ determined analytically equals 172.7 mm.

By using the nomogram in Fig. 2 it is possible to determine $\underline{m}$ and establish the value of the loss of pressure in the constricting device. For this purpose we find on the horizontal axis a division corresponding to C = 9.2. From

*In the examples only those physical parameters are given which are required for computations by means of nomograms.

this point we raise a perpendicular up to the curve representing the adopted pressure difference $h_{20\ max} = 630$ mm Hg. It will be seen from the nomogram that the required value of m is between 0.1 and 0.2. By interpolation it is possible to determine its value more accurately as 0.17. Along the vertical axis we determine the value of the loss of pressure $P_n = 0.73$ kg-wt/cm^2.

Example 2. The medium is natural gas consisting by weight of 61% methane, 10% ethane, 9% propane, 13% nitrogen, 1% butane, and other gases; the temperature of the medium is t = 5°C or T = 273 + 5 = 278°K; the excess pressure of the medium is $P_u = 2$ kg-wt/cm^2; the maximum flow is $Q_{n\ max} = 1600$ nm^3/hr; the internal diameter of the pipeline is $D_{20} = 104$ mm. The permissible loss of pressure is $P_{np} = 0.2$ kg-wt/cm^2.

The specific gravity of the dry mixture of gases of the above composition under normal conditions and calculated by the generally accepted method is equal to $\gamma_n = 0.953$ kg-wt/m^3. The total pressure of the gas mixture (medium) will be: $P = P_u + P_b = 2 + 1.033$ kg-wt/cm^2. The coefficient of compressibility of the mixture is K = 0.99.

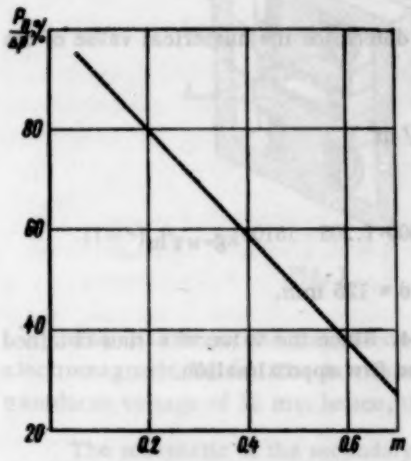

Fig. 3. Relation between the loss of pressure in a constricting device and m.

The gas flow, whose maximum value coincides with a standard value, is given in volumetric units. Hence we determine the auxiliary value C from (6), i.e.

$$C = (\alpha m\sqrt{\Delta P}) = \frac{Q_{n\,s\,max}\sqrt{\gamma_n TK}}{0.2108 D_{20}^2\sqrt{P}} =$$

$$= \frac{1600\sqrt{0.953.278.0.99}}{0.2108.10^4\sqrt{3.033}} = 6.52.$$

By moving vertically up from point C = 6.52 we obtain on the nomogram (Fig. 2) the possible value of $h_{20\ max}$. In doing so we come across the following points: $h_{20\ max} = 250$ mm Hg and $P_n = 0.29$, $h_{20\ max} = 160$ mm Hg and $P_n = 0.17$. The larger value of $H_{20\ max}$ raises the accuracy of measurements, but involves a value of P_n which exceeds considerably the permissible value of $P_{np} = 0.2$ kg-wt/cm^2. Hence we select the lower value of $h_{20\ max} = 160$ mm Hg, which corresponds to $P_n = 0.17$ kg-wt/cm^2.

In order to determine d from the nomogram of Fig. 1 it is first necessary to calculate the gravimetric flow $G_{max} = Q_n\gamma_n = 1600 \cdot 0.953 = 1525$ kg-wt/hr, the pressure differences Δp in kg-wt m^2, i.e., $\Delta p = 0.001\, h_{20}\gamma_{20} = 2170$ kg-wt/m^2 and the specific gravity of gas under operating conditions $\gamma = \gamma_n(P/T)\cdot(T_n/P_n) = 2.97$ kg-wt/m^3. For the first approximation we assume that $\varepsilon = 1$.

By using the nomogram in this manner we find that d is smaller than 100 mm. Hence we use the lower scale for the flow and divide by 10 the corresponding value of d obtained from the vertical scale. We then place the right-hand end of the strip against line D of a value 10 times greater than the given figure. For the above case we find that d = 49 mm.

Having determined by the normal method the correcting factor for the expansion of the measured medium ε [2] which is equal to 0.99, we find a new value of $G/\varepsilon = 1540$ kg-wt/hr. By means of this figure we determine more precisely the value of d which is d = 49.5 mm.

Example 3. The measured medium consists of the dry part of moist air; the maximum flow is $Q_{n\ max} = 12{,}000$ nm^3/hr; absolute air pressure is P = 7.033 kg-wt/cm^2; the temperature of the medium is t = 55°C or T = 328°K; air humidity is $\varphi = 0.5$; the permissible loss of pressure in the constricting device is $P_{np} = 0.2$ kg-wt/cm^2; the internal diameter of the pipeline is $D_{20} = 260$ mm.

The specific gravity of moist air calculated by the normal method is, under normal conditions, equal to $\gamma = 7.29$ kg-wt/m^3.

For the maximum flow we take a round figure equal to the nearest higher standard flow of $Q_{n\,s\,max} = 1.25 \cdot 10^3$ nm^3/hr.

The auxiliary quantity for humid air is determined, considering that $\varepsilon = 1$, from:

$$C=(\alpha m\sqrt{\Delta p})=\frac{Q_{n\,s\,max}\sqrt{\gamma}}{0.01252D_{20}^2}\cdot\frac{P_h}{P-\varphi P_{h\,max}}\cdot\frac{TK}{T_n}=\frac{12500\sqrt{7.29}\cdot1.033\cdot328\cdot1}{0.01252\cdot260^2(7.033-0.5\cdot0.161)273}=6.61.$$

We find from the nomogram in Fig. 2 that for the above value of C the most suitable values of other parameters are: $H_{20\,max} = 160$ mm Hg for which $m = 0.23$ and $P_n = 0.17$ kg-wt/cm^2; moreover, the last quantity does not exceed the maximum permissible value of the pressure loss in the constricting device $P_{np\,max}$ which is determined from the relationship

$$P_{np\,max}=P_{np}\frac{Q^2_{n\,s\,max}}{Q^2_{n\,max}}=0.2\left(\frac{12500}{12000}\right)^2=0.22\ \text{kg-wt/cm}^2$$

The next higher value of $h_{20\,max} = 250$ mm Hg cannot be taken, since it corresponds to $P_n = 0.28$ kg-wt/cm^2, which exceeds $P_{np\,max}$, thus failing to satisfy the given conditions.

In order to find the value of d from the nomogram in Fig. 1, we have to determine the numerical value of the pressure difference and the volumetric flow of air:

$$\Delta p=0.001h_{20}\gamma'_{20}=0.001\cdot160\cdot13546=2170\ \text{kg-wt/m}^2$$

$$G_{min}=0.01252\varepsilon\alpha mD_{20}^2\sqrt{\Delta p\gamma}=0.01252\varepsilon CD_{20}^2\sqrt{\gamma}\approx Q_{n\,s\,max}\quad\gamma_n=12500\cdot1.205=15100\ \text{kg-wt/hr}(\varepsilon=1).$$

By assuming that $\varepsilon = 1$ we determine from the nomogram in Fig. 1 that d = 125 mm.

By means of normal calculations we find a more exact value of $\varepsilon = 0.994$. Since the value of ε thus obtained differs very little from unity and the computation is only tentative we adopt the first approximation.

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ELECTROMAGNETIC FLOWMETER WITH ERRORS NOT EXCEEDING 0.5% OF THE FULL-SCALE DEFLECTION

L. M. Korsunskii and É. M. Madikyan

Translated from Izmeritel'naya Tekhnika, No. 8, pp. 53-55, August, 1961

The Khar'kov State Institute of Measures and Measuring Instruments has developed and produced an experimental model of an electromagnetic flowmeter with a top measuring limit of 500 m^3/hr, whose measuring error does not exceed 0.5% of its full-scale deflection.

The flowmeter consists of a transducer which generates an emf produced by the movement of the liquid in a magnetic field, and of a secondary measuring instrument.

The schematic of the transducer is shown in Fig. 1. Poles 1 of the electromagnet fed by a mains frequency current enclose a plexiglas chamber 2 with a rectangular channel passing through it. Two copper electrodes 3 are mounted inside the channel flush with its walls. The cross section area of the channel is 70 × 220 mm . The chamber is connected to the pipeline by means of junction pipes which provide a smooth contraction and expansion of the flow.

The electromagnet core and pole pieces are made of mixed transformer steel. A battery of capacitors which is connected in parallel with the magnet windings and tuned to the resonance frequency greatly reduces the current consumption. The power consumption of the instrument does not exceed 400 va.

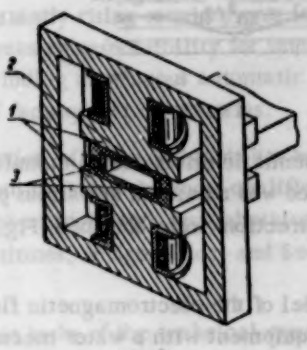

Fig. 1.

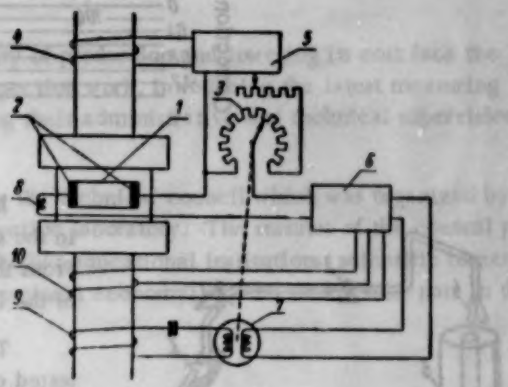

Fig. 2.

The chamber of the instrument is carefully screened in order to protect it from interference from extraneous electromagnetic fields. The interference level does not exceed 5 μv. A flow of 500 m^3/hr requires an output transducer voltage of 12 mv; hence, the ratio of the effective signal to interference is of the order to 2500.

The schematic of the secondary instrument is similar to that previously described [1], but it differs in certain details. The output voltage of the transistor is compensated by a voltage obtained from slide wire 3 (Fig. 2) where 1 is the chamber of the instrument and 2 its electrodes. The slide wire is fed from a special coil 4, placed on the joke of the instrument magnet. This prevents the variations of the magnetic field strength having any substantial effect on the instrument readings. In the previously described instrument the coil was placed in the gap of the electromagnet which led to considerable distortions of the magnetic field and additional errors.

The slide wire is fed through a special phase shift device 5. The difference voltage is amplified by a 7-tube RC coupled amplifier 6, which operates a reversible motor 7. The input impedance of the amplifier is of the order of 3 meg, which makes it possible to measure the flow of liquids whose resistivity does not exceed 10^5-10^6 ohm · cm.

It is known that one source of interference in electromagnetic flowmeters consists of the instrument's alternating magnetic field which induces in the measuring circuit a voltage not connected with the movement of the liquid. This voltage is shifted by 90° with respect to the signal. The reversible motor which balances the circuit is also used as a phase-sensitive element. Owing to this phase-sensitive element the secondary instrument does not react to the interference produced by the transducer's magnetic field. However, this interference overloads the amplifier, thus raising its threshold of sensitivity and making it necessary to use a rough form of compensation. Both the amplitude and phase must be compensated. With an inaccurate phase compensation a compensated voltage component will arise in the measuring circuit in phase with the measured signal and will be likely to produce considerable errors. The previously described tube compensator is sufficiently stable. Coil 8, consisting of several turns and capable of rotating about an axis perpendicular to the magnetic field, is connected in series with one electrode and is used for compensation purposes. The amplitude of the emf induced in the coil depends on its angle of rotation with respect to the magnetic field, and the phase of the emf is always displaced by ± 90° with respect to the signal.

In order to obtain stable phase relationships the excitation winding of the reversible motor and the anode circuits of the last amplifier stage are fed respectively from special windings 9 and 10, which are placed on the yoke of the electromagnet.

The kinematic circuits of normal electrode potentiometers are very complicated and produce additional errors. In this instrument the kinematic circuit has been greatly simplified. The casing of the slide wire and the reversible motor are fixed to a metallic plate which serves as a base for the instrument. The slide wire contact and the pointer are mounted on the same axis and firmly fixed with respect to each other. In order to reduce friction and strain the axle is mounted on ball bearings. The slide wire is protected from external pickups by means of a screen which also carries the scale of the instrument. The motor drives the slide wire axle through a gear transmission. Due to the rigid connection between the slide wire runner and the pointer, the secondary instrument kinematic circuit does not produce any additional errors.

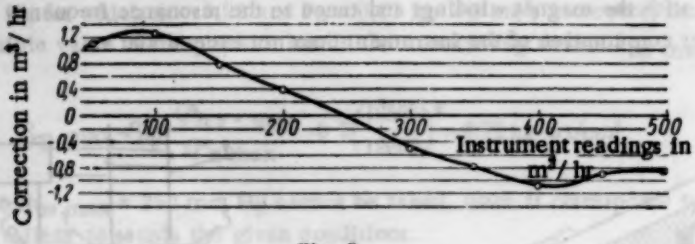

Fig. 3.

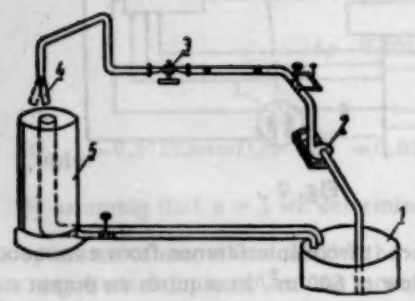

Fig. 4.

In order to find the systematic errors due to manufacturing defects in the slide wire, its resistance was measured at various points of the scale. From these measurements corrections were obtained (Fig. 3) for the scale of the instrument.

The experimental model of the electromagnetic flowmeter was tested on a volumetric test equipment with a water measuring tank of 27 cm^3 capacity. The layout of the equipment is shown in Fig. 4. Pump 2 draws the water from supply tank 1 and pumps it through the pipeline to the electromagnetic flowmeter 3. From the instrument the liquid discharges into measuring tank 5. A special switching device controls the flow of the liquid into the tank. Lever 4 of the switching device automatically connects and disconnects a stopwatch. The errors in estimating the time interval do not exceed 0.1-0.2 sec.

The measuring tank of the equipment was calibrated by means of a reference container which was certified with an error of 0.05%.

The accuracy of the electromagnetic flowmeter was evaluated by means of repeated measurements. At each measured point of the scale, three series of 7 measurements each were carried out. Each series was conducted on different days with intervals of 30-50 min between measurements.

The dispersion of readings in the mean of the three measured series is characterized by a mean square error of $\sigma = 0.4\text{-}0.5$ m^3/hr, which is 0.08-0.1% of the full-scale deflection.

If it is assumed that the maximum random error does not exceed 3σ, the possible errors of the instrument can be evaluated as

$$\delta = |\delta_0| + 3\sigma,$$

where δ_0 is the systematic error.

The systematic error is due in the main to inaccuracies in the manufacture of the slide wire and does not exceed 0.25% of the full-scale deflection. Thus, δ will not exceed 0.5% of the full-scale deflection.

In testing this instrument more than 180 measurements were made, and not a single measurement had an error exceeding 0.4% of the full-scale deflection, even without the application of corrections.

During measurements the electromagnet supply voltage was varied between 200-220 v without producing any noticeable effect on the instrument readings.

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MATERIAL RECEIVED BY THE EDITORIAL BOARD

EXPERIENCE OF THE WORK OF THE GKL (STATE INSPECTION LABORATORIES) TECHNICAL COUNCIL AND ITS SECTIONS*

Translated from Izmeritel'naya Tekhnika, No. 8, pp. 55 58, August, 1961

B. N. Vorontsov

The constantly rising requirements for improving the quality of production and lowering its cost face the GKLs with increasing responsibility for improving testing and inspection work, introducing the latest measuring equipment including continuous automatic testing, and improving their administrative and technical supervision of the activity of factory test laboratories.

The solution of these problems has been greatly assisted by the technical council which was organized by the Gor'kii GKL and included the most qualified workers of the inspection laboratory. The statutes of the council provide that members of factory test laboratories, the personnel of higher educational institutions, scientific research and design institutes, design offices and Sovnarkhoz (council of national economy) experts should take part in this work.

The main tasks of the technical councils consist of:

a) dealing with future plans for the development of laboratories and working out measures for their implementation;

b) dealing with the problem of introducing new measuring equipment and means of automatic control, and assisting the assimilation of this equipment at the plants and organizations of the Gor'kii economic area;

c) improving the supervision of plants and organizations which produce measuring equipment and checking the performance of newly produced measuring devices;

d) making suggestions to the Committee of Standards, Measures and Measuring Instruments on establishing standards for technical conditions in the production of various articles, and suggestions for amendments and additions to existing standards;

e) dealing with the normalization and unification of the production in basic and auxiliary industries, as well as introducing standards and norms at the plants of the economic area;

f) dealing with the most important rationalization suggestions for the improvement of measuring equipment and the perfection of the laboratory test equipment, as well as assisting the innovators and inventors in setting out their suggestions;

g) studying, unifying, spreading and helping to assimilate the experience of innovators in measurement techniques;

h) studying and preparing reports and conclusions on draft intructions and regulations of the Committee, as well as proposals made to the Committee for changing existing instructions and regulations;

i) improving the organization, planning and carrying out of inspection and checking;

j) the preparation and holding of seminars and technical conferences on the exchange of experience;

k) implementing the measures and recommendations approved by the technical council through existing sections of the Sovnarkhoz technical and economic council with the participation of representatives from the plants (heads of the measuring equipment service inspection agencies).

* The Editorial Board has received articles by B. N. Vorontsov and B. L. Sokolov on the experience of the work of technical councils in certain inspection laboratories for measuring equipment. These articles are published here.

The technical council has the right to:

a) entrust its members with developing the required material and preparing it for council sessions according to an approved plan;

b) involve in the working-out of certain problems engineers, technicians and other laboratory workers, as well as experts and innovators in the sphere of measurement techniques in plants and organizations of the Gor'kii economic area and representatives of the Sovnarkhoz;

c) convene conferences and organize teams for working out certain problems;

d) develop and submit for the consideration of the Sovnarkhoz technical and economic council administration recommendations on various problems coming within the competence of laboratories;

e) compile plans of work for the technical council and approve them;

f) organize, where necessary, trips for its personnel to other inspection and factory laboratories and other organizations, for the purpose of exchanging experiences.

At the beginning of the council's activity only a linear and angle measuring section was organized, and later sections for thermotechnical and electrical and radio measurements.

The first practical measures of the linear and angle measurement sections were directed toward intensifying the activity of the factory central test laboratories (TsIL). In this connection the section decided that each TsIL should compile its organizational and technical plan comprising the following sections:

1. Assimilation of new checking methods and equipment, mainly instruments and devices for production testing, designed to operate in the workshops.

The TsIL should pay particular attention to the assimilation of automatic control equipment and establishing the necessary conditions for its operation.

2. The improvement of the methods and operation of the measuring equipment service inspection agencies, which would provide a simplification and reduction of technical documents in checking, raising responsibilities in the use of the equipment, increasing its life, checking the measuring equipment in situ, developing simpler techniques for checking various instruments (checking the operating range of the scale), etc.

3. Study of the operational properties of new instruments. The service inspection agencies should continuously study the operational properties of newly received instruments and gauges, report their observations and suggestions to the manufacturing plants, and improve existing measures and measuring instruments.

4. Raising the qualifications of the personnel. Conduct technical instructions, seminars, sending people to courses for improving their qualification, exchanging experiences by visiting other plants for special purposes.

5. Improving the working conditions in the TsIL and the control and checking stations (KPP). Organization of work-benches according to the Committee's instructions, provision of normal temperature and lighting conditions, provisions of the basic and auxiliary equipment, instructions on testing especially complicated components and products, etc.

In compiling the plan of activity it is necessary to take into account the potentialities of the TsIL, the training of the personnel, etc. Such plans, compiled for half-yearly or yearly periods, are approved by the chief engineer or a person responsible for the uniformity of measurements at the plant, and constitute documents which oblige the TsIL to fulfil all the measures envisaged in the plan.

The supervision of the service inspection agencies by the inspection laboratory includes the checking of the fulfilment of these plans; moreover, the measures which deserve attention by the results of their application should be noted with approval in the inspection report.

At present the planning of organizational and technical measures has been firmly established in the routine work of the TsILs and has greatly intensified their activity. Of the various measures introduced by the service inspection agencies as a result of the operation of the linear and angle measuring section the following most effective measures should be noted:

a) a commission consisting of section members examined the service inspection personnel. Before the examinations technical instruction at the plants was organized;

b) a comparative checking of the work of the factory personnel was organized in the following types of measurements: checking grade 5 block gauges, checking thread calibers; checking smooth snap gauges, checking smooth rings by means of an optimeter, measuring hardness on a TK hardness gauge. The test results were discussed at the section with the participation of the persons who took part in the test checking;

c) technical conferences with demonstrations of new types of instruments were organized and proved to be of interest to many plants. As a result of these conferences the most efficient instruments and devices were introduced at several plants. For instance, instruments for measuring angle gauges (KPU-1) were produced and assimilated at 10 plants.

Instruments for checking external threads on bolts in situ were made and assimilated, Sofronov type projectors were acquired and introduced, drawings for making a device used in checking the screw-line curvature in thread calibers were drawn and distributed, and many devices developed by the GKLs and plant laboratories for checking instruments and gauges were produced;

d) three-months' courses for checking opticomechanical instruments were organized, seminars for the study of the principles of operation of continuous automatic testing instruments, and for determining and checking the helical line curvature in thread calibers were held, and a draft instruction for checking end gauges was explained;

e) the draft regulation for central measuring equipment laboratories, compiled by the Committee of Standards, Measures and Measuring Instruments, has been widely discussed. As a result of this one plant has already established a measurement techniques department, headed by the chief metrologist who reports directly to the chief engineer of the plant.

The Gor'kii GKL is planning to help the setting up in 1961 of measurement technique departments in several plants. As a result of the work of sections the activity of the service inspection agencies has been made considerably more efficient. In addition to their checking work these agencies are beginning to deal with the improvement of production processes which depend on measurement techniques. A constant exchange of experience between factory TsILs has been organized, and the ties between the inspection laboratories and the service inspection agencies have been strengthened.

The technical council of the inspection laboratory has been able to involve in its work the best metrologists of the plants, thus greatly enhancing its authority.

In order to implement most effectively the decisions of the technical council sections and make the most important decisions to be considered compulsory by the plants, the Gor'kii GKL has started in 1961 to develop the activity of the sections according to theplan of the technical and economic council of the Sovnarkhoz as one of the Council's permanent sections, in the manner provided in the Regulation on laboratory technical councils.

The greatest attention in the work of the technical council sections in 1961 will be paid to organizing measurement equipment central laboratories (TsLIT) at the plants, and to organizing courses and seminars for the purpose of raising the qualifications of service inspection and the Gor'kii GKL personnel.

The organization of TsLIT's will require time and a different approach to each plant, taking into consideration the peculiarities of their production and the condition of their service inspection agencies. It is impossible to reorganize services which have been in existence for many years simply by issuing orders to plants. It is necessary to show by convincing examples the efficiency and necessity of establishing TsLIT's as authoritative agencies capable of solving problems connected with production technical measurements.

B. L. Sokolov

Technical councils attached to the GKLs are being organized according to the resolution passed by the Russian Federation Republican Conference of the state inspection laboratories personnel, in September,1960.

The favorable experience gained from the operation of the Gor'kii GKL technical council is confirmed by the activity of technical councils of the Rostov, Smolensk, Tula, and other inspection laboratories. Thus, in connection with the establishment in the Smolensk Region of a new industry, namely instrument-making, the technical council of the Smolensk GKL discussed measures designed to prepare several instrument-making plants for the production of measuring instruments, the state testing of experimental batches of these instruments, and distribution of the technical documents involved in state testing. These conferences were attended by the GKL personnel, chief engineers and heads of the technical control divisions in the plants, representatives of the Sovnarkhoz engineering administrations of the Leningrad Special Design Office of the Academy of Sciences, USSR, of the Smolensk branch of the All-Union Scientific Research Institute of Thermal Instruments, etc.

Another equally important question discussed by the technical council was the organization of central test laboratories in the instrument-making plants of the Smolensk Sovnarkhoz.

The technical council of the Tula GKL, together with the Sovnarkhoz technical department, received reports from administrators of four plants on the introduction of standards, republican technical specifications, industrial and service norms at the plants, and examined proposals made in this connection.

Available experience in the operation of technical councils shows that a most decisive factor for the success of their activity consists in the composition of the council.

It is very important to include in the council leaders of the GKL measurement group, outstanding workers and representatives of social organizations, experts from the Sovnarkhoz, and other service organizations, leading workers from plants, factory test laboratories, control and measuring-instrument workshops, pioneers and innovators in measurement techniques, and others. For a more expert approach to the problems envisaged by the development plan it is advisable to involve in the work of the technical council members of scientific research and design institutes, design offices, experts in measurement techniques and others.

The technical council must be guided in its activity by the Regulation approved by the Plenipotentiary representative of the Committee attached to the Executive Committee of the local Soviet, namely, the head of the GKL. The technical council may form sections for various types of measurements.

The Council is headed by a chairman, who is elected at an open session of the council and conducts his work according to a quarterly plan.

General meetings of the council are held once a quarter, meetings of industrial sections at least once a quarter. The technical council and its sections conduct their work in close collaboration with the production and technical councils of plants, establishments and organizations. If necessary the meetings of the sections can be held at plants or institutes.

The preparation of reports and draft resolutions of the technical council and its sections is made by the members of the council on instructions from the council or its sections as well as by representatives or plants or organizations by special instruction of the council or its sections. The reports and draft resolutions to be discussed at the council meeting must be submitted to the secretary of the council at least 5-7 days prior to the meeting, and those for discussion at the section meetings to the section leaders 5-7 days before the meeting. Decisions of the technical council and its sections are recorded in the minutes and approved by the Committee's Representative at the Executive Committee of the local Soviet.

The experience of work of technical councils organized by some GKLs has shown the present and future importance of their activity. It is necessary to support and develop in every way the establishment of the GKLs' technical councils.

IMPORTANT TASKS OF GKL (STATE INSPECTION LABORATORY) WORKERS IN AGRICULTURE

V. V. Petropavlovskii

Translated from Izmeritel'naya Tekhnika, No. 8, pp. 58-59, August, 1961

In order to implement the decisions of the January (1961) plenary session of the CPSU (Communist Party of the Soviet Union) central committee for eliminating important shortcomings in determining the quantity and quality of grain and technical crops, the fat content in milk, and other qualitative indexes, the personnel of state inspection laboratories (GKLs) are faced with the responsibility for a considerable improvement of their work in helping to assimilate new, modern measuring equipment in all branches of agriculture, checking the correct use of measuring equipment and, in the first place, the equipment used for determining the quantity and quality of agricultural products.

State inspectors know how to check scales, hondrometers, moisture meters and fat content meters, but this is insufficient. Since they operate in rural districts where agricultural products are being delivered, they must be able, if required, to make check weighing of grain, determine the degree of its purity, its type and humidity, as well as determine the fat content of milk. At the same time the state inspectors must check the ability of stocking organizations personnel to use correctly their measuring equipment and, if required, provide them with the necessary instruction. They must constantly and effectively assist the newly organized agencies of state inspection of purchases to determine the quality and weight of agricultural products, providing the required advice and instructions; they should strive to provide the newly organized model farms with up-to-date measuring equipment, and see to it that the personnel of these farms are able to carry out correctly the required qualitative and quantitative measurements.

In order to enable the GKL personnel to immediately tackle these important problems in agriculture it is necessary to organize appropriate technical instruction for them, using as leaders for such courses qualified workers in laboratories, elevators, dairy plants and establishments of local stocking organizations. In many cases it will be advisable to hold such courses on the premises of these organizations, making use of their equipment.

In acquiring the technique for determining the quality and moisture content of grain and the fat content of milk, the state inspectors must not only learn how to carry out these tests, but they must perform them with confidence and efficiently, meeting fully and accurately the appropriate state standards as well as service instructions and specifications. Special attention should be paid to the technique of taking samples, since the correct determination of the above qualitative indexes depends, to a great extent, precisely on the methods of selecting test samples. Grain samples should be taken according to GOST (All-Union State Standard) 3040-55, and milk samples according to GOST 3622-57.

In order to direct this new activity of the GKLs toward a definite goal, it is advisable to instruct several laboratories to check-test on a mass scale the qualitiative indexes of grain and milk, and to present the material thus collected to the Committee of Standards, Measures and Measuring Instruments, for its classification and analysis. This would provide the possibility of ascertaining the reasons for incorrect measurements. It would then become possible on the basis of such an analysis to adopt the required organizational and technical measures aimed at a radical improvement in determining the qualitative indexes of grain and milk.

SELECTION OF METHODS AND MEANS OF MEASUREMENT IN PRODUCTION

A. A. Gaikov and N. D. Belyi

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The wide development of cooperation in industrial production sets important requirements with respect to the interchangeability of the produced commodities. In providing such conditions of interchangeability, the measurement technology becomes a most important factor.

The aim of keeping within set tolerances, i.e., of making interchangeable the commodities produced not only at a certain time in a given plant, but also those produced at any time in any of our plants,can only be achieved by the appropriate organization of measurements.

The system of tolerances at present adopted in engineering satisfies in the main the interchangeability requirements. However, effective interchangeability cannot be provided even with a perfect system of tolerances, unless this system is accompanied by a strict uniformity and correct application of measures and measuring instruments in various plants of our country. Any neglect in observing uniform measures has often led, and leads, to difficulties in production cooperation.

In this connection it is most important to correctly choose the appropriate equipment for checking technological processes according to the type of production, accuracy of machining, the degree of mass-production and other production characteristics. In each particular case it should be decided which of the technical inspection methods should be adopted: Whether it should be static, continuous automatic, automatic or semiautomatic inspection of finished products, inspection by means of measuring devices, etc. It should be noted that inspection automation as a rule not only makes the work of inspector much easier, but makes production checking less expensive though more objective and reliable.

A reduction in the cost of acceptance inspection can be attained, without lowering the quality of production, by prophylactic checking, i.e., by studying and eliminating the causes of defects. It often happens in practice that a new automatic checking machine exposes many more defects than had been the case before its use. This is due to the fact that a correctly operating automatic machine brings to light the so-called hidden defects, thus indicating the necessity of improving the technological process. Its improvement may make one hundred per cent checking and the use of the automatic machine no longer necessary. It may be found, however, that the former inspection methods provide a sufficiently high quality of production. In this case it is not advisable to change the technological process and the checking machine need not be used in this instance either.

The measuring methods and equipment are selected on the basis of their economic factors, which include the cost of the equipment, duration of its operation before resetting is required, the time required for setting it to a given dimension, the time involved in the measuring process, and the training required for the servicing personnel.

The Tashkent GKL checked the manner in which some plants selected measuring methods and equipment, and found that crude mistakes were still being made in the planning of technological processes. Thus, the Tashkent gauge plant issued an instruction for testing the tapering and the difference in the diameters of the working part of a end-type milling cutter with an odd number of teeth by means of a micrometer. Such a measurement, however, is impossible since these cutters have not a single pair of teeth diametrically opposite each other. These parameters can only be measured, for instance, on a universal microscope type UIM-21.

The Tashkent cable plant also made an incorrect selection in certain production methods and means of measurement. At all the testing stations the resistance of conductors below 2 ohm was measured on single bridges which, owing to their constructional peculiarities, have large errors in measuring small resistances. This parameter should have been measured on a double bridge which has an error not exceeding 1.5%.

In the same plant ac cable-testing sets use grade 2.5 instead of grade 0.5 instruments, which does not provide the required accuracy of measurements. Insulated conductors were tested on an equipment which had no continuous voltage control and did not provide a voltage with a stability of ± 2.0% during testing, its voltage variations

amounted to ± 0.5%. The above equipment was not provided with means for measuring the test voltage, and as a result some of the tests were carried out with a voltage 3.0-39.0% below the specified value.

The yarn impregnating installation in the power cable shop was fitted with 25 kg-wt/ cm^2 manometers, whereas the maximum measured operating pressure amounted to 3 kg-wt/ cm^2, thus the impregnation was not measured with sufficient accuracy.

An incorrect choice of measuring equipment distorts the technological process and affects the quality of production. The fact that certain production, for instance, power cables of the Tashkent cable plant, was revealed by inspection to be below the required standards is not accidental.

It is just as important in high-quality production that the measuring equipment should be maintained at the required level, i.e., it must always be usable and accurate. It is often found in practice that, although the methods and means of measurement have been correctly selected, their defective condition results in the production being below the requirements set by state standards and specifications, in a waste of electricity and raw materials, in defective computations of electrical energy, steam, water, etc.

It is known that the maintenance of measuring equipment at the required level of efficiency is entrusted to service inspection agencies, consisting of factory (TsIL, TsZL, TsLIT, KPP, etc) [central test laboratory, central factory laboratory, central measuring equipment laboratory, control and checking stations, etc] and area laboratories for measuring equipment. The latter service plants which have only a small number of instruments. State inspection laboratories of measuring equipment exercise continuous supervision of plant and area laboratories.

On the basis of the decisions of the July (1960) plenary session of the CPSU (Communist Party of the Soviet Union) central committee, many inspection service agencies of the Uzbek SSR Sovnarkhoz (Council of national economy) have reorganized their work and are now paying greater attention not only to testing, but also to introducing improved methods and measuring equipment and to studying the operational qualities of the instruments in use. Guided in their activity by the service inspection regulations of the ministry or department to which their plants belong, they are now taking part in umpire testing of the quality of production and in other questions connected with measuring equipment. In addition to providing testing graphs of the instruments at present in use, the above agencies have developed future plans for introducing new measuring equipment, as well as plans for studying the operational properties of measuring instruments, and have organized the testing of instruments in their laboratories.

The above work has already resulted in considerable savings for the plants, due to the improvement of technological processes, reduction of rejects, increased labor productivity, etc. Thus, the Uzbekenergo area test laboratory has produced a potentiometric equipment with a semiautomatic potentiometer R2/1, which has raised the labor productivity of the inspectors by 15-20%.

For testing insulation of low-voltage motors and other equipment of Chirchik electrochemical combine has introduced a 500 v megger with an electronically stabilized voltage. Only one man is required to operate this instrument instead of the two normally required for other types of meggers.

The replacement of uneven-arm 0.5 ton scales by automatic scales at the Nukus will has provided a saving in wages of 1700 rubles per year; the use of a 100-ton weighbridge at the Dzhambaisk stocking station has resulted in the saving of 48,000 rubles per year and in addition has reduced the standstill time and raised the turnover of railroad cars.

However, besides the service inspection agencies which have reorganized their work there are still some agencies which work in the inefficient "old way." These agencies do not introduce new measurement methods or equipment, the condition of the existing equipment in the plants under their supervision is often unsatisfactory. In this connection the Tashkent GKL has been forced to introduced, in addition to other measures, an over-all heavy state inspection of all the measures and measuring instruments at several of these plants.

Such inspections have a decisive effect on the condition of the measuring equipment in the plants which fall below the required standards, and forces them to organize systematic supervision of their measuring equipment. However, the wide application of these measures is undesirable, since it distracts inspection laboratories from their many other organizational and technical tasks.

In order to maintain the existing measuring equipment at the required level of efficiency, introduce more effective and better measuring methods and equipment, study the operational properties of existing measures and measuring instruments for the purpose of the further development of our instrument-making industry, the service inspection agencies should speed up the reorganization of their work.

INSTRUMENT TESTING SPECIFICATIONS SHOULD BE DIFFERENTIATED

K. V. Ovchinnikov

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p. 61, August, 1961

The existing specified outputs for grade 1, 2, and 3 GKIs (State inspection laboratories) place the grade 2 and 3 laboratories, which do most of the instrument checking, in a very difficult position, owing to the differences in their provision with reference instruments, and hence in the different methods of checking they are compelled to use. The specifications provide that the following numbers of reference instruments should be tested per man-shift: 7 resistance boxes, 7 bridges or dc potentiometers with at least 4 significant figures, 5 ammeters or voltmeters types M80/1 and M82 with 20 measuring ranges each, etc. These production norms can be attained or exceeded only by working overtime, which may have detrimental effects on the quality of the work.

We consider that the production norms should be differentiated according to the grade of the GKI without reducing the Committee's over-all plan of work.

It also seems advisable to simplify the forms used for reporting test results. For instance, in checking resistance boxes type KMS-6 several hundred measurements are made according to specification. It is not necessary to record the errors of all these measurements, it is perfectly satisfactory to enter only the maximum errors for each decade, i.e., 6 entries in all, thus saving a considerable amount of time. The same may be said with respect to reference pointer instruments, in which for each scale only one maximum correction should be calculated, instead of 10 corrections for ammeters and 15 for voltmeters, as it is done at present.

Finally, it is desirable to increase the exchange of experience between various GKIs and to apply everywhere the better methods adopted by some of the laboratories.

INFORMATION

BRITISH TRADE FAIR

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The British Trade Fair was held in the Moscow Sokol'niki Park from May 19 to June 4, 1961.

A considerable part of the Fair was occupied by various measuring instruments used in research laboratories and in production control and automation. The most comprehensive displays consisted of electrical, electronic and radio instruments. Some of the exhibited instruments had already been shown at last year's British Laboratory and Industrial Instruments Exhibition.*

Blackburn Electronics exhibited a five-digit transistorized voltmeter type BIE 2116. The instrument has a projection type display and an automatic indication of polarity, and measuring range between 2 and 1000 v. Its error is ±0.01%, sensitivity 10 μv, transformation time 90 msec, and input impedance 20 meg. It is provided with an output for a printing device. It weighs 12 kg and its power consumption is 200 w.

The transistorized type BIE 362 dc amplifier has a gain adjustable in the range of 10 to 300, its linearity is 0.1%, zero drift referred to the input of 3 μv in 8 hr, bandwidth up to 25 kc, input impedance of 3 meg, and its weight is 8 kg.

The Solartron Company exhibited digital display voltmeters, digital converters for feeding information to computers, and an ac converter voltmeter.

It is interesting to note this firm's dc digital voltmeter type LM 902.2 with a measuring range of 0.1 mv-1599 v with an error of 0.1%, and the ac converter voltmeter type LM 903 with a conversion range of 0.01 to 159.9 v for frequencies of 20 cps to 5 Mc. This converter, in conjunction with instrument LM 902.2 provides a digital display.

The Cambridge Company's electronic recording potentiometer type 74151 has a high-speed operation (0.4 sec) and a good damping. The instrument has no velocity feedback and uses a motor with a large starting torque. It measures up to 1 mv with an error not exceeding 0.5% and a sensitivity of at least 1 μv. A six-channel model is produced with a printing speed of 50 results per minute.

The same firm exhibited a multichannel boiler water analyzer. It is noteworthy that in this analyzer the same secondary instrument records concentrations of oxygen, hydrogen, hydrogen ions and electrical conductance.

Of considerable interest is an electronic potentiometer with three measuring ranges (1, 5 and 10 mv) and a facility for discrete displacement of zero in 22 steps of 1, 5, and 10 mv each, depending on the measuring range. In all the potentiometers reference elements consist of silicon stabilitrons. The potentiometers of this firm have an oil-immersed slide wire with rubbin contacts.

The Kent Company's electronomic potentiometer Mark 3 Multelec has an error of 0.25% and a sensitivity of 0.1%. It measures up to 2 mv. Its power pack uses Zener diodes. The instrument weights 64 kg.

The Marconi Instrument Company exhibited an AM and FM signal generator, a sweep generator, double-beam oscilloscope, pulse generators, signal shape analyzer, a power meter, ac microvoltmeters, tube millivoltmeters and voltmeters, a Q-meter, and universal bridges.

The Marconi electronic counter type TF 1165 has a capacity of 9999999 and can count pulses of an amplitude of 0.5 to 30 v with a frequency up to 1.0 Mc. Its input resistance is 1 meg, input capacity 50 μμf. It weighs 23 kg and has a neon-type digital display. In conjunction with timer type TF 1220 it can measure frequency and oscillation periods.

The Honeywell Company exhibited commercial electronic potentiometers with a reading settling time of 0.25 sec and a measuring circuit power pack with Zener diodes, flowmeters, liquid level and pressure indicators, and miniature pneumatic instruments.

* Izmeritel'naya Tekhnika 1960, No. 8.

The Pye Company's Dynacap pH meters have 9 ranges of (pH) 0-10, 4-14, 0-2, 2-4, 4-6, 6-8, 8-10, 10-12, and 12-14. The smallest calibration of all the ranges with the exception of the first is equal to 0.02 pH. The drift per 24 hr does not exceed ±0.03 pH. Its input resistance is 10^{13} ohm. The same firm exhibited galvanometers, a galvanometer amplifier, a precision decade potentiometers, a dc amplifier, an argon chromatograph, and an infrared spectrophotometer.

The Cossor Company exhibited double and single-beam miniature oscilloscopes and a number of radio-measuring instruments.

The Daw Company showed bridges, potentiometers, audio and ultrasonic generators, and stroboscopes.

The Cintel Company showed a rectangular pulse generator with a range from 1 cps to 1 Mc, a duty factor of 2 and an output voltage of 50 v. The rise time is $10 \cdot 10^{-9}$ sec, the distortion of the flat top of the pulse does not exceed 1%, the frequency is controlled with an accuracy of ± 2%. The counter and universal timing device shown by the firm are transistorized, use printed circuits, and indicating pointer instruments.

The firm EIL exhibited a millisecond interval meter type 25 E which can measure time intervals of 0.2 msec to 1 sec with an error of ± 2%. The instrument has an input impedance of about 2 meg and responds to negative and positive pulses of 3 to 300 v.

The electronic millivoltmeter with a dynamic capacitor type 33 B made by the same firm has the following characteristics: measuring range of 10.0 to 1000 mv, sensitivity of at least 0.1 mv, zero drift not exceeding 0.1 mv per 12 hr. Input impedance of 10^{13} ohm.

The EIL Company also exhibited portable and commercial pH meters, titrators, fluxmeters and megamegaohmmeters.

The leading manufacturers of equipment for measuring flow and pressure of liquids and gases were well represented at the Exhibition.

The De Havilland Aircraft Company exhibited a set of turbine flowmeters of the Pottermeter type from 3 to 100 mm in diameter with a top capacity of 30 liter/ hr to 540 m^3/ hr and a measuring range of 20 : 1. Their error was ±0.5%. Operating temperature −270 to +820°C. To special order the firm makes instruments to measure flows up to 10,000 m^3/ hr.

The Honeywell Controls Company exhibited a grade 0.5 bellows differential manometer for working at temperatures from −40° to +120°C and a pneumatic converter of pressure differences based on force compensation (weight 10.5 kg).

The Kent Company exhibited a wide range of instruments including mercury float differential flowmeters with a disc diaphragm (driven by an 8-day clock mechanism), and an integrator, bellows differential manometers for measuring pressure, flow, level, density, etc. Instruments equipped with self-protecting bellows, devices for damping pulsations, thermal compensators for temperatures between −32° and + 71°C, auxiliary, devices for measuring temperatures between −40° and +600°C. It also exhibited bell-type differential manometers with a top measuring range of 12 to 200 mm H_2O equipped with three-pen recorders. Static pressure meters up to 3.5 kg-wt/ cm^2, aircraft turbine flowmeters with miniature transducers, weighing 312 g and having top measuring limits of 0.9 to 4.5 m^3/ hr. Their measurement error is (in the range 1 : 10) ±1%. Another version of these meters is produced with a manual correction of fuel density variations in the range of 0.68-0.86 g/ cm^3. Their measuring range is 22.7-454 kg/ min or 22.7-227 kg/ min. Their error is (in the range 10 : 1) ±1.5%. Their weight is 765 g. Doll's tube has an extremely low residual pressure loss (lower than that of a Venturi tube) for a short length (half that of a Venturi tube).

The Meterflow Company's turbine flowmeters type M1 with their indicating and integrating secondary instruments are of interest. Each instrument is calibrated separately and supplied with a calibration curve. The nonlinearity of their range of 8 : 1 amounts to ±0.25%. Their operating temperature is from −180° to +150°C, a pressure of 210 kg-wt/ cm^2 (by special order the firm makes transducers for temperatures up to 450°C and higher pressures). The measuring range is from 1.6-13.6 to 145-1182 liter/ min. The same firm produces for the oil industry turbine flowmeters type M5 with ranges from 136-1364 liter/ min to 0.9-9 m^3/ hr. The instability of their readings is smaller than 0.1% of the full-scale deflection. The loss of pressure at the maximum flow amounts to 0.2 kg-wt/ cm^2. The operating temperature is −200 to +300°C. The firm also makes miniature induction pressure converters (including

those of an absolute type) with a top measuring limit of 420 kg-wt/cm^2. Their weight does not exceed 255 g.

The Armstrong Whitworth Equipment Company exhibited a miniature liquid flow speed transducer which consists of a 5-vane turbine whose axis is fixed at the end of the tube in such a manner that the vanes pass in front of the tube end at a distance of 0.1 mm from a gold wire placed inside the tube. The wire is connected to a bridge circuit whose output voltage is a measure of the flow speed. Its measuring range is from 23 mm/sec to 0.6 m/sec. Its error amounts to ±5% at a speed of 25 mm/sec, ±2% at 755 mm/sec, and ±1% at 150 mm/sec and over. Its dimensions are 15 × 25 × 15 mm. The firm also exhibited miniature high-pressure flowmeters (up to 420 kg-wt/cm^2) with a constant pressure difference and a range from 2.25-22.5 to 13.5 liter/min and an error of ±2% of the measured value in the range of 1 : 4 and ±5% over the first quarter of the scale. Their operating temperature is from 0 to 100°C.

The Langham Thompson Company exhibited a set of miniature pressure converters suitable both for static and dynamic measurements. The UP and BP type converters use strain gauges. Their top measuring limits range from 250 mm H_2O to 1400 kg-wt/cm^2, their error is between ±1.5 and ±0.5% of the full scale deflection respectively. Their operating temperature extends from −40° to +120°C. Their weight is between 200 and 300 g.

Bellows transducers intended for measuring pressure differences up to 1.4 kg-wt/cm^2 at pressures up to 21 kg-wt/cm^2 and down to 7600 mm Hg. Their permissible overload amounts to 100%. Their error is ±1.5%. Their weight is 680-960 g. Transducers whose sensing elements consist of a single-turn tubular spring are designed for top-scale pressures from 3.5 to 210 kg-wt/cm^2. Their error is ±1.5% of the full scale deflection. Their weight is 300 g.

The Negretti and Zambra firm exhibited manometers with a single turn tubular spring (measurement range of 0-4.2 to 0-700 kg-wt/cm^2, measurement error of ±0.5% for top limits of 70 kg-wt/cm^2 and 1% for the remaining ranges); manometers with a diaphragm unit (top measuring limit up to 2 kg-wt/cm^2), aircraft flowmeters with a constant pressure difference (indicating and integrating) and a measurement error of ±0.7% of the full scale deflection and a total error of integration of ±2% in the range of 1 : 5, their measuring range extends from 4.5-90 to 4.5-180 kg/min; pneumatic pressure transducers based on the force balancing principle.

The Cambridge Instrument Company exhibited a set of indicating manometers and vacuum meters with metallic and nommetallic (slack) diaphragms (top measuring limit up to 150 mm H_2O), with single-turn tubular springs (top limit up to 700 kg-wt/cm^2), Pitot tubes, recording manometers (with the same range as indicating manometers), electrical pressure transducers.

The automatic measuring system for dispensing liquid oil products exhibited by the Lindars Automation Company is worth noting. The dispensing can be accomplished by volume or by weight. The quantity of the measured-out liquid is automatically printed by means of a summing device or an electromagnetically operated typewriter.

SEMINAR FOR IMPROVING QUALIFICATIONS

G. I. Klykov

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p. 63, August, 1961

The seminar for improving the qualifications of the TsIL and KPP (Central test laboratory and Control and checking stations) personnel at engineering plants, organized by the Central Bureau of Technical Information of the Leningrad Sovnarkhoz (Council of national economy) at the M. I. Kalinin Leningrad Polytechnical Institute has completed its work.

The 105 students at the seminar included heads of TsIL and BTKs (Technical inspection bureaus), TsIL engineers, foremen, TsIL and KPP laboratory assistants and inspectors of the largest engineering plants in Leningrad.

Lectures on the following subjects were given at the seminar: 1) design and methods of checking the equipment for measuring dimensions in engineering—80 hr; 2) tolerances and fits in engineering—28 hr; 3) geometrical optics—48 hr.

The seminar rendered considerable assistance to the production workers in becoming familiar with modern measurement techniques and brought to their notice the latest standards and instructions of the Committee of Standards, Measures and Measuring Instruments.

After completing their studies at the seminar the students continue to receive advice and attend instructional excursions to plants. Thus the TsBTI of LSNKh (Central bureau of technical information of the Leningrad Sovnarkhoz) organized an excursion to the Leningrad instrument plant "LIZ" where the students were shown the manufacture of components, assembly and testing of the manufactured measuring equipment. Excursions were also organized to other plants of the city. Twenty of the best students were rewarded with a trip to the Moscow instrument plants "Kalibr" and MIZ, where they studied the production technique and exchanged experiences.

The students wish to express their gratitude to the organizers of the seminar, and particularly to engineer A. V. Mel'nikov of the TsBTI of LSNKh and to the lecturers of the M. I. Kalinin Leningrad Polytechnical Institute who lectured at the seminar: I. S. Amosov, the chief of the laboratory A. L. Markov, and V. N. Dorina.

COMMITTEE OF STANDARDS, MEASURES AND MEASURING INSTRUMENTS

Translated from Izmeritel'naya Tekhnika, No. 8,
p. 64, August, 1961

I. New Specifications for Measures and Measuring Instruments Approved by the Committee

New Standards (Registered in May and June, 1961)

GOST (All-Union State Standard) 4381-61. Lever micrometers. Replacing GOST 4381-57. Effective from July 1, 1962.

GOST 7328-61. Weights. Replacing GOST 7328-55. Effective from January 1, 1962.

GOST 9763-61. Radio measuring instruments. General technical requirements. Effective from October 1, 1961.

GOST 9771-61. Heterodyne frequency meters. Technical requirements. Effective from October 1, 1961.

GOST 9772-61. Resonance frequency meters. Technical requirements. Replacing GOST 10191. Effective from October 1, 1961.

GOST 9776-61. Instruments for checking worms, worm gears and worm gear transmissions. General technical requirements. Effective from January 1, 1962.

GOST 9781-61. Radio-measuring instruments. Electronic voltmeters. General technical requirements. Effective from January 1, 1962.

GOST 9788-61. Standard signal generators. List of parameters. Effective from January 1, 1962.

Operating Instructions Issued for Checking Measures and Measuring Instruments and Approved by the Committee's Measuring Instruments Administration

Operating instruction No. 179 for checking surface thermocouples.

Operating instruction No. 180 for checking reference loaded-piston vacuum meters type MVP-2.5 Order II (Grade 0.05) with mutually balancing pistons.

Operating instruction No. 181 for checking reference loaded-piston scales type OGV-1 for a range of 0-1000 kg.

II. Measures and Measuring Instruments Approved by the Committee as the Result of State Tests, and Passed for Use in the USSR (Registered in April and May, 1961)

Three-phase moving-iron phasemeters, trade mark É771 and É772, of the Krasnodar Sovnarkhoz (Council of national economy). State Register No. 1454-61.

Moving-iron ammeters and voltmeters, trade mark É761 and É762, of the Krasnodar Sovnarkhoz. State Register No. 1455-61.

Ferrodynamic frequency meters, trade mark D761 and D762, of the Krasnodar Sovnarkhoz. State Register No. 1456-61.

Fixed attenuators, trade mark FAD-100/30, of the Leningrad Sovnarkhoz. State Register No. 1458-61.

Fixed attenuators, trade mark FAD-5/20, FAD-5/30, of the Leningrad Sovnarkhoz. State Register No. 1459-61.

Fixed attenuators, trade mark FAD-5/80, of the Leningrad Sovnarkhoz. State Register No. 1460-61.

Mobile 2-ton platform scales with a dial indicator, trade mark VPTs-2 (m), of the Krasnodar Sovnarkhoz, State Register No. 1461-61.

Automatic rationing unequal-arm scales, trade mark DS-800, of the Kiev Sovnarkhoz. State Register No. 1462-61.

Tensile-strength-testing machines for measuring samples with an effort of 500 kg-wt, trade mark MR-0.5, of the Ivanovo Sovnarkhoz. State Register No. 1463-61.

Tensile-strength-testing machines for measuring samples with an effort of 50 kg-wt, trade mark MR-0.05, of the Ivanovo Sovnarkhoz. State Register No. 1464-61.

Mercury-in-glass thermometers for gas analyzers type UG-1, of the Poltava Sovnarkhoz. State Register No. 1465-61.

Mercury-in-glass thermometers for refractometers RL, of the Poltava Sovnarkhoz. State Register No. 1466-61.

Mercury-in-glass thermometers for hygrometers type VCh, of the Poltava Sovnarkhoz. State Register No. 1467-61.

Graduated test tubes with normal slides with a capacity of 5, 10, 15, and 20 ml, of the Moscow regional Sovnarkhoz. State Register No. 1468-61.

Portable cable-testing instrument, trade mark PKP-2 (m) of the Ministry of Communications, USSR, entered in the State Register under No. 731-61, replacing cable-testing instrument KP entered in the State Register under No. 731.

Rack-mounted ammeters and voltmeters, trade mark Ts362,.of the Krasnodar Sovnarkhoz. State Register No. 1469-61.

Portable voltamperewattmeters, trade mark D129, of the Leningrad Sovnarkhoz. State Register No. 1470-61.

Portable moving-coil voltammeters, shock- and vibration-proof, trade mark M128, of the Leningrad Sovnarkhoz. State Register No. 1471-61.

Combined voltmeters, trade mark M369, of the Krasnodar Sovnarkhoz. State Register No. 1472-61.

Soviet Journals Available in Cover-to-Cover Translation

ABBREVIATION	RUSSIAN TITLE	TITLE OF TRANSLATION	PUBLISHER	TRANSLATION BEGAN Vol.	Issue	Year
AÉ	Atomnaya énergiya	Soviet Journal of Atomic Energy	Consultants Bureau	1	1	1956
Akust. zh.	Akusticheskii zhurnal	Soviet Physics – Acoustics	American Institute of Physics	1	1	1955
	Antibiotiki	Antibiotics	Consultants Bureau	4	1	1959
Astr(on). zh(urn).	Astronomicheskii zhurnal	Soviet Astronomy–AJ	American Institute of Physics	34	1	1957
Avto(mat). svarka	Avtomaticheskaya svarka	Automatic Welding	British Welding Research Association (London)		1	1959
	Avtomatika i Telemekhanika	Automation and Remote Control	Instrument Society of America	27	1	1956
	Biofizika	Biophysics	National Institutes of Health*		1	1957
	Biokhimiya	Biochemistry	Consultants Bureau	21	1	1956
Byull. éksp(erim). biol. i med.	Byulleten' éksperimental'noi biologii i meditsiny	Bulletin of Experimental Biology and Medicine	Consultants Bureau	41	1	1959
DAN (SSSR) / Dokl(ady) AN SSSR	Doklady Akademii Nauk SSSR	The translation of this journal is published in sections, as follows:				
		Doklady Biochemistry Section (Life Sciences)	American Institute of Biological Sciences	106	1	1956
		Doklady Biological Sciences Sections (Includes: Anatomy, biophysics, cytology, ecology, embryology, endocrinology, evolutionary morphology, genetics, histology, hydrobiology microbiology, morphology, parasitology, physiology, zoology sections) (Life Sciences)	American Institute of Biological Sciences	112	1	1957
		Doklady Botanical Sciences Sections (Includes: Botany, phytopathology, plant anatomy, plant ecology, plant embryology, plant physiology, plant morphology sections) (Life Sciences)	American Institute of Biological Sciences	112	1	1957
		Proceedings of the Academy of Sciences of the USSR, Section: Chemical Technology (Chemical Sciences)	Consultants Bureau	106	1	1956
		Proceedings of the Academy of Sciences of the USSR, Section: Chemistry (Chemical Sciences)	Consultants Bureau	106	1	1956
		Proceedings of the Academy of Sciences of the USSR, Section: Physical Chemistry (Chemical Sciences)	Consultants Bureau	112	1	1957
		Doklady Earth Sciences Sections (Includes: Geochemistry, geology, geophysics, hydrogeology, mineralogy, paleontology, petrography, permafrost sections) (Earth Sciences)	American Geological Institute	124	1	1959
		Proceedings of the Academy of Sciences of the USSR, Section: Geochemistry (Earth Sciences)	Consultants Bureau	106-123	1 6	1957-1958
		Proceedings of the Academy of Sciences of the USSR, Sections: Geology (Earth Sciences)	Consultants Bureau	106-123	1 6	1957-1958
		Doklady Soviet Mathematics (Mathematics)	The American Mathematics Society	131	1	1961
		Soviet Physics–Doklady (includes: Aerodynamics, astronomy, crystallography, cybernetics and control theory, electrical engineering, energetics, fluid mechanics, heat engineering, hydraulics, mathematical physics, mechanics, physics, technical physics, theory of elasticity sections) (Physics)	American Institute of Physics	106	1	1956
		Proceedings of the Academy of Sciences of the USSR, Applied Physics Sections (does not include mathematical physics or physics sections) (Physics)	Consultants Bureau	106-117	1	1956-1957
Derevoobrabat. prom-st'.	Derevoobrabatyvayushchaya promyshlennost'	Wood Processing Industry	Timber Development Association (London)		9	1959
	Élecktrosvyaz	Telecommunications	Massachusetts Institute of Technology*		1	1957
Éntom(ol). oboz(renie)	Éntomologicheskoe obozrenie	Entomological Review	American Institute of Biological Sciences	38	1	1959
Farmakol. (i) toksikol(ogiya)	Farmakologiya i toksikologiya	Pharmacology and Toxicology	Consultants Bureau	20	1	1957
FMM	Fizika metallov i metallovedenie	Physics of Metals and Metallography	Acta Metallurgica*	5	1	1957
Fiziol. zhurn. SSSR (im. Sechenova)	Fiziologicheskii zhurnal im. I. M. Sechenova	Sechenov Physiological Journal USSR	National Institutes of Health*		1	1957
Fiziol(ogiya) rast.	Fiziologiya rastenii	Plant Physiology	American Institute of Biological Sciences	4	1	1957
	Geokhimiya	Geochemistry	The Geochemical Society		1	1958
FTT	Fizika tverdogo tela	Soviet Physics–Solid State	American Institute of Physics	1	1	1959
Izmerit. tekh(nika)	Izmeritel'naya tekhnika	Measurement Techniques	Instrument Society of America		1	1959
Izv. AN SSSR, O(td). Kh(im). N(auk)	Izvestiya Akademii Nauk SSSR: Otdelenie khimicheskikh nauk	Bulletin of the Academy of Sciences of the USSR: Division of Chemical Sciences	Consultants Bureau		1	1952

continued

Izv. AN SSSR, O(td). T(ekhn). N(auk): Met(all). i top.	(see Met. i top.)				
Izv. AN SSSR Ser. fiz(ich).	Izvestiya Akademii Nauk SSSR: Seriya fizicheskaya	Bulletin of the Academy of Sciences of the USSR: Physical Series	Columbia Technical Translations	1	1954
Izv. AN SSSR Ser. geofiz.	Izvestiya Akademii Nauk SSSR: Seriya geofizicheskaya	Bulletin (Izvestiya) of the Academy of Sciences USSR: Geophysics Series	American Geophysical Union	1	1954
Izv. AN SSSR Ser. geol.	Izvestiya Akademii Nauk SSSR: Seriya geologicheskaya	Izvestiya of the Academy of Sciences of the USSR: Geologic Series	American Geological Institute	1	1958
Kauch. i rez.	Kauchuk i rezina	Soviet Rubber Technology	Research Association of British Rubber Manufacturers	18	1 1959
	Kinetika i kataliz	Kinetics and Catalysis	Consultants Bureau	1	3 1960
	Koks i khimiya	Coke and Chemistry USSR	Coal Tar Research Association (Leeds, England)		1 1958
Kolloidn. zh(urn).	Kolloidnyi zhurnal	Colloid Journal	Consultants Bureau	14	1 1952
	Kristallografiya	Soviet Physics – Crystallography	American Institute of Physics	2	1 1957
Metalov. i term. obrabot. metal.	Metallovedenie i termicheskaya obrabotka metallov	Metal Science and Heat Treatment of Metals	Acta Metallurgica	6	1 1958
	Metallurg	Metallurgist	Acta Metallurgica		1 1957
Met. i top.	Metallurgiya i topliva	Russian Metallurgy and Fuels	Eagle Technical Publications		1 1960
Mikrobiol.	Mikrobiologiya	Microbiology	American Institute of Biological Sciences	26	1 1957
OS	Optika i spektroskopiya	Optics and Spectroscopy	American Institute of Physics		6 1959
	Pochvovedenie	Soviet Soil Science	American Institute of Biological Sciences		1 1958
	Priborostroenie	Instrument Construction	British Scientific Instrument Research Association		1 1959
Pribory i tekhn. éks(perimenta)	Pribory i tekhnika éksperimenta	Instruments and Experimental Techniques	Instrument Society of America		1 1957
Prikl. matem. i mekh.	Prikladnaya matematika i mekhanika	Applied Mathematics and Mechanics	American Society of Mechanical Engineers		1 1958
PTÉ	(see Pribory i tekhn. éks.)				
	Problemy Severa	Problems of the North	National Research Council of Canada		
Radiotekh.	Radiotekhnika	Radio Engineering	Massachusetts Institute of Technology*	12	1 1957
Radiotekh. i élektronika	Radiotekhnika i élektronika	Radio Engineering and Electronics	Massachusetts Institute of Technology*	2	1 1957
	Stanki i instrument	Machines and Tooling	Production Engineering Research Assoc.		1 1959
	Stal'	Stal (in English)	Iron and Steel Institute		1 1959
Stek. i keram.	Steklo i keramika	Glass and Ceramics	Consultants Bureau	13	1 1956
Svaroch. proiz-vo	Svarochnoe proizvodstvo	Welding Production	British Welding Research Association		4 1959
Teor. veroyat. i prim.	Teoriya veroyatnostei i ee primenenie	Theory of Probability and Its Applications	Society for Industrial and Applied Mathematics		1 1956
Tsvet. Metally	Tsvetnye metally	Nonferrous Metals	Primary Sources		1 1960
UFN	Uspekhi fizicheskikh Nauk	Soviet Physics – Uspekhi (partial translation)	American Institute of Physics	66	1 1958
UKh	Uspekhi khimii	Russian Chemical Reviews	The Chemical Society (London)		1 1960
UMN	Uspekhi matematicheskikh nauk	Russian Mathematical Surveys	London Mathematical Society	15	1 1960
Usp. fiz. nauk	(see UFN)				
Usp. khim(ii)	(see UKh)				
Usp. matem. nauk	(see UMN)				
Usp. sovr. biol.	Uspekhi sovremennoi biologii	Russian Review of Biology	Oliver and Boyd	48	1959
Vest. mashinostroeniya	Vestnik mashinostroeniya	Russian Engineering Journal	Production Engineering Research Assoc.	4	1959
Vop. gem. i per. krovi	Voprosy gematologii i perelivaniya krovi	Problems of Hematology and Blood Transfusion	National Institutes of Health*	1	1957
Vop. onk.	Voprosy onkologii	Problems of Oncology	National Institutes of Health*	1	1957
Vop. virusol.	Voprosy virusologii	Problems of Virology	National Institutes of Health*	1	1957
Zav(odsk). lab(oratoriya)	Zavodskaya laboratoriya	Industrial Laboratory	Instrument Society of America	25	1 1959
ZhAKh Zh. anal(it). khimii	Zhurnal analiticheskoi khimii	Journal of Analytical Chemistry USSR	Consultants Bureau	7	1 1952
ZhETF Zh. éksperim. i teor. fiz.	Zhurnal éksperimental'noi i teoreticheskoi fiziki	Soviet Physics–JETP	American Institute of Physics	28	1 1955
ZhFKh Zh. fiz. khimii	Zhurnal fizicheskoi khimii	Russian Journal of Physical Chemistry	The Chemical Society (London)		7 1959
ZhMÉI Zh(urn). mikrobiol. épidemiol. i immunobiol.	Zhurnal mikrobiologii, épidemiologii i immunobiologii	Journal of Microbiology, Epidemiology and Immunobiology	National Institutes of Health*		1 1957
ZhNKh Zh(urn). neorgan(ich). khim(ii)	Zhurnal neorganicheskoi khimii	The Russian Journal of Inorganic Chemistry	The Chemical Society (London)		1 1959
ZhOKh Zh(urn). obshch(ei) khimii	Zhurnal obshchei khimii	Journal of General Chemistry USSR	Consultants Bureau	19	1 1949
ZhPKh Zh(urn). prikl. khimii	Zhurnal prikladnoi khimii	Journal of Applied Chemistry USSR	Consultants Bureau	23	1 1950
ZhSKh Zh(urn). strukt. khimii	Zhurnal strukturnoi khimii	Journal of Structural Chemistry	Consultants Bureau	1	1 1960
ZhTF Zh(urn). tekhn. fiz.	Zhurnal teknicheskoi fiziki	Soviet Physics–Technical Physics	American Institute of Physics	26	1 1956
Zh(urn). vyssh. nervn. deyat. (im. Pavlova)	Zhurnal vysshei nervnoi deyatel'nosti (im. I. P. Pavlova)	Pavlov Journal of Higher Nervous Activity	National Institutes of Health*		1 1958

*Sponsoring organization. Translation through 1960 issues is a publication of Pergamon Press.

SIGNIFICANCE OF ABBREVIATIONS MOST FREQUENTLY ENCOUNTERED IN SOVIET TECHNICAL PERIODICALS

AN SSSR	*Academy of Sciences, USSR*
FIAN	*Physics Institute, Academy of Sciences USSR*
GITI	*State Scientific and Technical Press*
GITTL	*State Press for Technical and Theoretical Literature*
GOI	*State Optical Institute*
GONTI	*State United Scientific and Technical Press*
Gosénergoizdat	*State Power Press*
Gosfizkhimizdat	*State Physical Chemistry Press*
Goskhimizdat	*State Chemistry Press*
GOST	*All-Union State Standard*
Gostekhizdat	*State Technical Press*
GTTI	*State Technical and Theoretical Press*
IAT	*Institute of Automation and Remote Control*
IF KhI	*Institute of Physical Chemistry Research*
IFP	*Institute of Physical Problems*
IL	*Foreign Literature Press*
IPF	*Institute of Applied Physics*
IPM	*Institute of Applied Mathematics*
IREA	*Institute of Chemical Reagents*
ISN (Izd. Sov. Nauk)	*Soviet Science Press*
IYap	*Institute of Nuclear Studies*
Izd	*Press (publishing house)*
LÉTI	*Leningrad Electrotechnical Institute*
LFTI	*Leningrad Institute of Physics and Technology*
LIM	*Leningrad Institute of Metals*
LITMiO	*Leningrad Institute of Precision Instruments and Optics*
Mashgiz	*State Scientific-Technical Press for Machine Construction Literature*
MGU	*Moscow State University*
Metallurgizdat	*Metallurgy Press*
MOPI	*Moscow Regional Pedagogical Institute*
NIAFIZ	*Scientific Research Association for Physics*
NIFI	*Scientific Research Institute of Physics*
NIIMM	*Scientific Research Institute of Mathematics and Mechanics*
NIKFI	*Scientific Institute of Motion Picture Photography*
NKTM	*People's Commissariat of the Heavy Machinery Industry*
Obrongiz	*State Press of the Defense Industry*
OIYaI	*Joint Institute of Nuclear Studies*
ONTI	*United Scientific and Technical Press*
OTI	*Division of Technical Information*
OTN	*Division of Technical Science*
RIAN	*Radium Institute, Academy of Sciences of the USSR*
SPB	*All-Union Special Planning Office*
Stroiizdat	*Construction Press*
URALFTI	*Ural Institute of Physics and Technology*
TsNIITMASh	*Central Scientific Research Institute of Technology and Machinery*
VNIIM	*All-Union Scientific Research Institute of Metrology*

NOTE: Abbreviations not on this list and not explained in the translation have been transliterated, no further information about their significance being available to us — *Publisher.*

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